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ANNUAL CONSTRUCTION ISSUE

NEW HAM RADIO

In this issue...

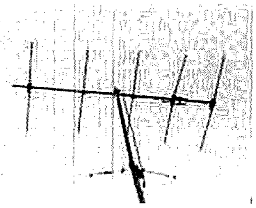
Computer Aided Design of Printed Circuit Boards

Making Printed Circuit Boards

The QRP TLC-Keyer

75-Hz Audio Filter

HAM RADIO



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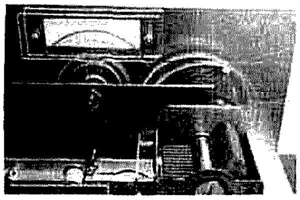

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Special thanks to NY1H for letting us use his equipment.

FEATURES

DEPARTMENTS

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Backscatter

The FCC and 20-Meter Third Party Traffic

Now hold on. Let me get this straight.

Hams are using the airwaves for business communications?

Can't be. We're better'n the rest and police ourselves.

Don't we?

All the latest hubbub about phone patches on 20 meters is very interesting. For years, many of us have wondered and agonized over what constitutes "business communications." Trying to be very careful, I have always followed the rule: if there's a doubt in my mind, don't do it.

A few years back the FCC waded into a big bruhaha about 2-meter autopatches and what was and was not permissible. It always struck me as silly that I couldn't call my wife on the way home to see if anything was needed for dinner. All I'd get in return is a better dinner. There's no profit or gain for me in the call. But it was one of those questionable calls and, consequently, not on the approved list.

The FCC has now decided to get involved in the morass of 20-meter third party traffic. I have the feeling that it may be too little too late. If you haven't listened to the top end of 20 lately, you have missed some interesting conversations.

According to W5YI, Robert McNamara, chief of the FCC Special Services Division, sent a letter to many of the participants active in third party traffic. The gist of McNamara's letter is that Amateur Radio should be for experimentation purposes in the field of communications, technical advancement, and the fostering of international good will. However, he says, "It appears that as much as 15 percent of the 20-meter [Amateur] band is being claimed for such communications [phone patch traffic] by some Amateur operators and by organizations apparently organized specifically to utilize the amateur service for third party telephony communications." W5YI says that the FCC has also received letters from other countries complaining about American Amateurs and their phone patch traffic. McNamara also states that "requests to the commission to resolve the dispute are taking far too much staff time from other essential activities."

Perhaps if the FCC had shown some leadership in the first place, they wouldn't have to solve the problem now.

This complaint has been voiced by many over the past ten years and is all too pertinent now. While I think we all agree that deregulation has worked to improve the overall quality of Amateur Radio, there are a number of areas that have slipped too far for even the most liberal to accept. From malicious interference, to foul language and poor operating habits, the total lack of enforcement has fostered this kind of behavior.

Self-policing only works when all parties involved agree to participate. Get one guy who refuses to go along and the system has been compromised. Unfortunately, there are too many "iconoclasts" who refuse to cooperate in a self-policing environment.

And so, you have a mess like the one we have on 20 meters. It's a shame and not in the true Amateur spirit. Hopefully, the FCC will take a strong leadership position to stop the abuses before they lead to serious damage to Amateur Radio. We have enough threats to worry about as it is without adding third part traffic violations to the primordial soup. We do not recommend a return to the restrictive policies of the FCC of forty years ago. However, the FCC is the only organization that has the power and ability to stop the abuses now, before irreparable harm is done.

It's kind of tough to extol the virtues of Amateur Radio to a friend and when you go to an on-the-air demonstration run across two bums arguing. You couldn't sell me on the basis of a demonstration like that. No sirree.

Craig Clark, NX1G

Editor's Notes

A Preview of Coming Attractions

Every year, we at *Ham Radio* try to bring you the best of Amateur Radio. Last year we introduced a redesigned editorial package that included new graphics, two Weekender projects per month, more construction articles, and lots of ham notes. As we enter the nineties, *Ham Radio* has two new features to offer you. Here are the details...

A lot of you have written regarding the status of W1JR's "VHF/UHF World" column. After five years of writing interesting and insightful articles about the upper regions for *Ham Radio*, Joe Reisert is taking a well-deserved break. After a long search, we are pleased to announce that Bob Atkins, KA1GT, former author of the "New Frontier" column in *QST*, has agreed to share his knowledge and enthusiam with you. We'll be bringing you his new column "Microwaves" as soon as the first installment arrives. Of course, Joe is still a much valued member of the *Ham Radio* family and we hope to present an occasional piece by him in the future. In the meantime, Joe's excited that Bob will writing for all you VHF/UHFers out there!

Also new is the "*Ham Radio* Bulletin Board." The "Bulletin Board" will feature some of the more technical *letters to the editor* that we receive — letters that go into too much detail to be used in our "Comments" section. It will appear in the magazine whenever we have messages for you, or you have messages for us.

I hope you enjoy these two new features. Both are part of our continuing effort to keep *Ham Radio* fresh, exciting, and full of the most up-to-date technical information available.

Terry Northup, KA1STC

HAM RADIO READER EVALUATION

It's been over a year since we introduced our editorial redesign of *Ham Radio* Magazine. At your request we've included more short construction projects, two weekender projects each month, and loads of ham notes. HR's 25-percent increase in circulation over the last 12 months must mean that you really like what you see in the "New" *Ham Radio*!

But we're not going to stop and rest on our laurels. Here's another opportunity to let us know how you feel about Amateur Radio's #1 technical magazine. Please take a moment to fill out and return this questionnaire — a photocopy is fine. Let us know how you feel about the "New" *Ham Radio* Magazine.

1. Do you like the changes we've made to *Ham Radio*?
   Yes ______ No ______ Why or why not? ______________________________
2. What else would you like to see? More technical articles ______ More projects ______
   More Weekenders ______ Other ______________________________
3. Do you like the current mix of articles? Yes ______ No ______
   Why or why not ______________________________
4. Do you think *Ham Radio* is: too technical ______, not technical enough ______, just right ______.
5. I like the Old ______ New ______ *Ham Radio* better. I don't really have a preference. ______
   Why? ______________________________

Please return your questionnaires to *Ham Radio* Reader Evaluation, Ham Radio Magazine, Main Street, Greenville, New Hampshire 03048. Thanks!

Comments

QSL cards available

Dear HR

The American Amateur Radio Club of Korea has in hand 55 pounds of QSL cards, some as much as 10 years old. These were recently released to KARL by the Postal Department. If you were (or know someone who was) a United States Forces Korea "HL9" during the past 10 years, a SASE to: American Amateur Radio Club of Korea, P.O. Box 153, APO San Francisco, California 96206, will get your cards. Don't forget to include your HL9 callsign and the dates it was valid.

Charles H. Kelley, W5SPK/HL9CK, PSC Box 5349, APO San Francisco, California 96366-0006.

Balance of Modes

Dear HR

Donald Sinex is right ("Comments," September). Morse code is no longer an indispensable skill for ham radio operating, and the inability to master it should not disqualify a person from becoming a ham. Code virtuosity — like typing speed — is not an indication of general intelligence.

Code stands or falls as a technical requirement. But license exams still overemphasize code even though other modes are predominant. This has led hams to view code as a fraternity initiation. "I did it, so you're going to have to do it too!"

We don't need fraternity initiations. CW is no longer the only mode we have and our license exams should reflect a *fair* balance of modes, instead of being 75 percent Morse.

Care to argue? See you on 21.060.

Michael A. Covington, N4TMI, Athens Georgia

New FCC rules

Dear HR

I just read the new September 1989 FCC part 97 rules for ham radio. It states on page 22 that "repeaters can be limited to certain user stations" when in a ancillary capacity, such as a parade or special function.

In checking the 1989 ARRL repeater directory for the 440-MHz band, I found 293 "closed to the licensed public" repeaters in the Southern California (Los Angeles) area. I could find only three "open to the public" repeaters. That means 99 percent of the Southern Californi repeaters are not available to the licensed public!

I feel we should notify the public to use the 440 repeaters, or possibly the 440 repeater band in Southern California, as a "test location" for the new "code-free" licensed public!

W. Douglass, K6BAZ, Glendale, California

True believer

Dear HR

Congratulations on your editorial in the September 1989 issue of *Ham Radio*. You have presented a balanced, concise viewpoint which is also well written.

While I was not especially keen on your magazine, you are making a believer out of me! Recent issues have been very informative, and I now look forward each month to receiving my copy.

Keep up the good work.

Marvin J. Fein, Newport Beach, California

Celebrity info request

Dear HR

I am researching a book on the life, times, and activities of Arthur Godfrey. In addition to his broadcasting, he was very active in many other areas: Amateur Radio, aviation, military service, stage shows, horsemanship, ecology, etc. I understand Mr. Godfrey had a very early, unusual callsign. Can anyone tell me what it was?

I may be reached at (201) 386-1920. Thank you for any ideas or reminiscences you or others may be able to share with me.

Lee R. Munsick, Whippany, New Jersey

Godfrey's callsign was K4LIB. Ed.

Article stirs memories

Dear HR

It has been a long time since I've enjoyed an article more than the one in the November issue on the early Hallicrafters' receivers. It was well done in every respect. I commend the author for writing it and you for publishing it. It's an example of the kind of "balance" we're seeing in the "new" *Ham Radio*.

My own affinity for Hallicrafters' products began in a similar fashion. I started my ham career just after World War II with the much-maligned, but eminently affordable, S-38. It surely offered very little in electronic sophistication, but because it covered 10 meters, it opened a new door on life for me! Those first years after the war found 10 meters wide open, often till well after dark, with hams all over the globe getting back on the air and signals rolling in from the far-flung corners. I was just an SWL back then, but what joy that little S-38 brought me.

After a year or so, I decided it was time to "trade up." My next receiver

(continued on page 76)

COMPUTER AIDED DESIGN
OF PRINTED CIRCUIT BOARDS

By Bryan P. Bergeron, NU1N, 30 Gardner Road, Apt. 1G, Brookline, Massachusetts 02146

Construction has always been one of the challenges of Amateur Radio. Before there were transistors and integrated circuits, the basic building blocks for most projects included a few tube sockets, a metal chassis, a handful of insulated terminals, and some wire, along with a few tubes, capacitors, resistors, and inductors. Today, the printed circuit board is used universally in electronic construction. The neatness and miniaturization made possible by etched pc boards, coupled with the additional benefits of low lead inductance, good physical stability, and repeatability, make it an ideal construction medium for group projects. With the aid of a personal computer and printer, you can design pc board etching patterns for use with the photographic circuit board preparation process.

Photographic reproduction

Photographic reproduction is the most efficient and error free means of transferring a layout from a printed page to a circuit board. Many journal articles and reference books feature pc board layouts. Photographic reproduction starts with the creation of a negative from the circuit layout artwork. A negative is usually created by copying, with a plain paper copier, the circuit board artwork onto transparent sheets used for overhead projections. This negative is pressed against a specially prepared photosensitive copper-clad board and exposed to ultraviolet (UV) light to deactivate the etchant resist on the exposed part of the board. You can use inexpensive UV exposure lights[1] or sunlight to expose the board. After removing the exposed copper with a standard etching solution, you can drill the board and solder the components into place.

One method of creating circuit board artwork is to transfer rub-on resist patterns to a white, nonporous surface. Once designed, the resist pattern can be used with a plain paper copier to create the required photographic negative. This method of circuit board design has some limitations. First, minor changes in artwork may require that you scrap the original design (and the press-on materials). Second, it's difficult to determine component placement, because you don't have a perfboard-like template to work with. Third, once a pattern is developed, it can't be used within another circuit board design easily. A computer-based drawing tool and a good quality dot matrix or laser printer are more efficient means of creating reusable circuit board artwork.

Computer hardware

You can use any computer capable of driving a dot matrix or laser printer (including the IBM PC and clones, Commodore, Amiga, and Macintosh) to generate circuit board artwork. Although a high resolution video display and a mouse, joystick, or other pointing device can make the task of creating circuit board artwork more enjoyable, the equipment you use is much less important than the quality of the printer output and the software tools available.

Output devices

The quality of your pc board artwork is to a large degree dependent on the quality of the printer or plotter used to print the tracing. A clean, sharp original provides the basis for a crisp circuit board tracing. A good quality laser printer,

like the LaserWriter (Apple Computer) for the Macintosh and the LaserJet (Hewlett-Packard) for IBM PC and compatibles, produces the best output. Even the smallest details of a tracing design can be reproduced clearly at 300 dots per inch (DPI). One of the advantages of using a laser printer is that you can create the photographic negative in one step, by loading the plain paper tray with heat-resistant transparency film. Film suitable for use in a plain paper copier will work.* On the down side, laser printers are expensive; prices fall in the $1000 to $3000 range.

The dot matrix printer is suitable for reproducing moderately complex circuit board tracings. The quality and design of your printer defines the limits of tracing density and complexity. Because most dot matrix printers for home use have resolution of near 100 DPI, the output won't be as clear, the curves won't be as smooth, and the optical density of the printer output won't be as great as compared with the output available from a laser printer. Unevenness in the optical density of dot matrix printer output, caused by variations in the printer ribbon, can be minimized with a new ribbon. The process of copying the artwork to transparency film will also minimize variations in the optical density. Another way to reduce these variations is to prepare an enlarged version of your circuit and shoot it down to size. This method is popular with professionals who have access to lithographic equipment. You can use a reducing copier in the same way; however, there's a tradeoff involved. There will be some loss of quality with an office copier — especially if it's not well maintained. You'll need to calibrate out an "indeterminate" reproduction ratio.

Plotters can also be used to create a hard copy of circuit board artwork. While comparable in price to low-end laser printers, small plotters offer little if any advantage over a laser printer. But plotters do excel in the ability they give you to handle circuit board tracing designs larger than those that can be accommodated by the standard paper sizes supported by most laser printers.

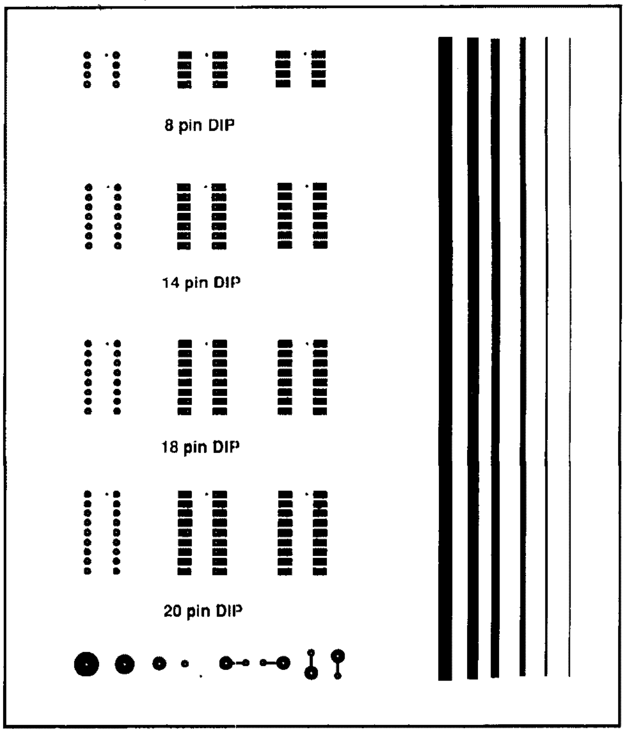
Software

Although programs are available that will compute automatically the trace patterns required to connect a given group of components in a multilevel board (e.g., HiWire for the PC from Wintek Corporation), the price of these programs can equal or exceed that of a good HF or VHF transceiver. A more economical alternative is to use one of the popular drawing programs to design the trace patterns. Once drawn to scale, a circuit board pattern can be sent to a printer to create the photo resist pattern.

There are two basic types of drawing programs available for most microcomputers. The first, typified by MicroSoft Paintbrush (MicroSoft Corp.) for the IBM PC and MacPaint (Claris Corp.) for the Apple Macintosh, can produce free-form graphics with great variability in shading, color, and contour. These "bitmap" editors produce graphic documents in a manner analogous to using wet paint on a canvas. Once individual graphics have been created and combined, there's no easy way to recover the individual elements. Free-form bitmap editors, while great for drawing lifelike scenes of people and places, lack many of the features needed to support the design of pc board traces.

The second type of drawing programs, object-oriented

FIGURE 1



A subset of a library of foil pattern objects constructed in an object-oriented graphics editor (MacDraw). This example illustrates foil patterns for 8 to 20 pin integrated circuits (left), component and connector patterns (bottom), and interconnecting tracings of various widths (right). Objects from this and other libraries are fashioned into a circuit board by copying and pasting the patterns onto a worksheet (see *Figure* 3).

graphics editors, produce documents composed of separate reusable objects. Even if it's combined with a line or other graphic object, a circle remains a circle — just as if you were working with cardboard cutouts. Object-oriented graphic programs — typified by MacDraw (Claris Corp.), Cricket Graph (Cricket Software), and SuperPaint (Silicon Beach Software) for the Apple Macintosh, and Windows "DRAW!" (Micrografx, Inc.), Freelance Plus (Lotus Development Corp.), and AutoCAD (AutoDesk, Inc.) for the IBM PC — are ideally suited to creating pc board patterns.* Objects can be grouped into other objects, and these new objects can be moved or copied to other areas on the working document. For example, a group of suitably arranged circles or rectangles can be grouped into IC tracings. The ability to save predefined objects in a library makes for quick and easy circuit board design (see Figure 1).

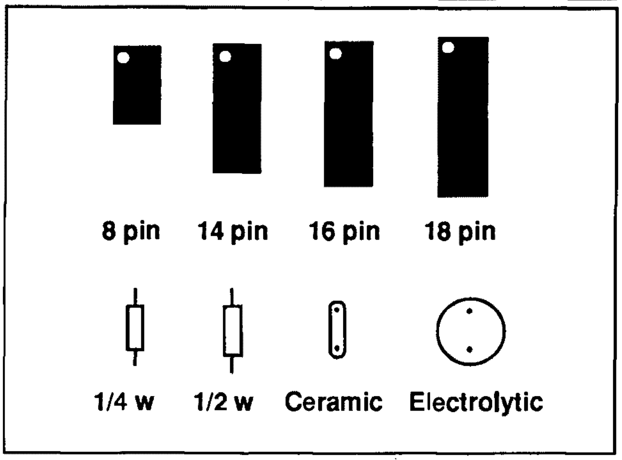
An example system

The system I use to draw printed circuit board artwork is based on the Apple Macintosh MacDraw software and the Apple LaserWriter printer. Like the better object-oriented drafting programs available for the Amiga, IBM PC, or Commodore 64, MacDraw supports the concepts of objects, object groups, and independent layers. As discussed earlier, objects can be grouped together to form other

*Try TEC Film, available from The Meadowlake Corporation, Box 497, Northport, New York 11768.

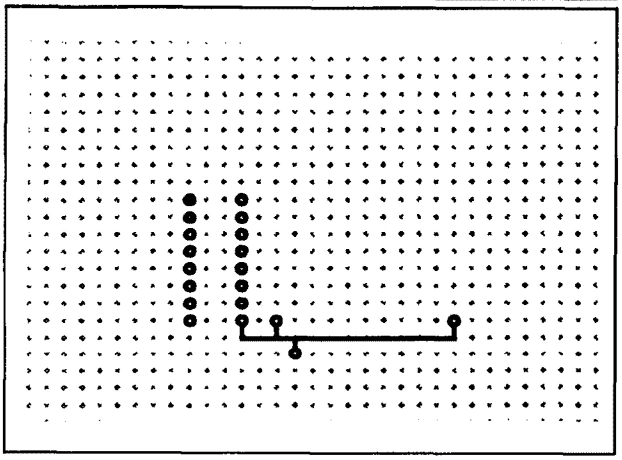
*These programs and others are available from most computer stores and software dealers.

FIGURE 2



Examples from a component library constructed in an object-oriented drawing program (MacDraw). Standard integrated circuit (top), resistor (left, bottom), and capacitor (right, bottom) bodies can be arranged and rearranged on a worksheet until the physical layout of the circuit board is adequate. Component objects are not only useful in illustrating circuit component layout, but they can be used to determine the minimum distance between circuit board tracings.

FIGURE 3



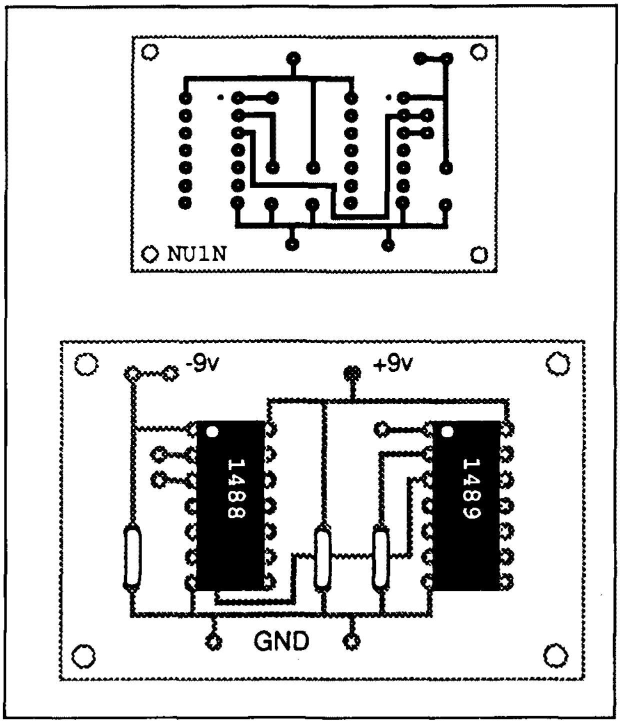
The worksheet, manipulated as a layer by the object-oriented graphics program. Much like the layout paper sold by the Vector Company, the computer-based worksheet serves as a template for the layout of a circuit board. In this illustration, the foil pattern for a 16-pin IC has been added, as well as a few connection points to pin 8 of the IC.

objects, which can be copied easily and moved to working documents (see **Figures** 1 and 2).

Figure 3 illustrates the layers concept. The background layer, which serves as both a guide and a template, is composed of a matrix of dots with a spacing and size identical to that of standard perfboard. Graphic objects (see **Figure 1**) can be copied and placed over the simulated perfboard layer to create a circuit board pattern. Layers can be hidden from view and printed selectively. For example, the perfboard guide in **Figure 3** can be hidden in the final printed output (see **Figure 4**, top). Other layers can be added to the printed output (see **Figure 4**, bottom) to show component placement.

Figure 4 also illustrates some of the powerful functions associated with object-oriented graphics programs — the ability to scale, flip, rotate, or invert objects easily. The top picture in **Figure 4** is actual size as viewed from the foil side; the bottom picture is viewed from the component side at 200-percent actual size.

FIGURE 4



The completed, actual size circuit board tracing design (top), along with the component side design at 200 percent actual size (bottom). Notice that the worksheet pattern does not appear in the final output. This board, together with three 0.1 μF capacitors and the two ICs indicated, provides for bi-directional TTL to RS-232 conversion.

Conclusion

I've used the system described here to develop circuit board patterns for projects as small as the one illustrated in **Figure 4**, and for boards containing more than a dozen integrated circuits. In general, I've found that the time savings afforded by the computerized approach is directly proportional to the complexity of the project. Massive alterations in a tracing pattern can be made in a matter of minutes instead of hours. If your club is interested in promoting construction projects, you should seriously consider borrowing or acquiring a system like the one described above. You'll never want to work with tape and rub-on resist patterns again.

REFERENCES

1. L. Hala and P. Hala, "UV Exposure Light," *Modern Electronics*, March 1989, page 28-34.

75-HZ WIDE
AUDIO FILTER

IC header-type construction makes this easy to build

By Albert A. Roehm, W2OBJ, 22 Brookdale Road, Cranford, New Jersey 07016

I recently worked on developing a tactile pad for the hearing impaired. The project required an evaluation of several different kinds of audio filters. I explored Gaussian versus Butterworth filters, envelope detection, time delay circuits to absorb noise pulses, and carrier-activated limiters. Eventually, I decided that simpler was better, and used a bandpass amplifier and comparator to drive some digital circuits. Later, I extracted and repackaged the filter portion of the project as an effective audio filter for some of the local hams. Why not add this filter to your receiving setup? The parts are readily available and the IC header-type construction makes it easy to complete the project in an evening or two.

Circuit description

Designers of good receivers have always tried to obtain the ultimate in selectivity. The migration from single-tuned L-C circuits, to cascaded IF stages, to crystal and mechanical filters gradually approached the desired rectangular, or 1:1, shape factor. But it was necessary to balance the losses of the selective circuits against the danger of instability that resulted when too much gain was used to overcome these losses. The designers also had to be careful to avoid ringing in passive or active audio filters, especially where the response curve contained a sharp peak.

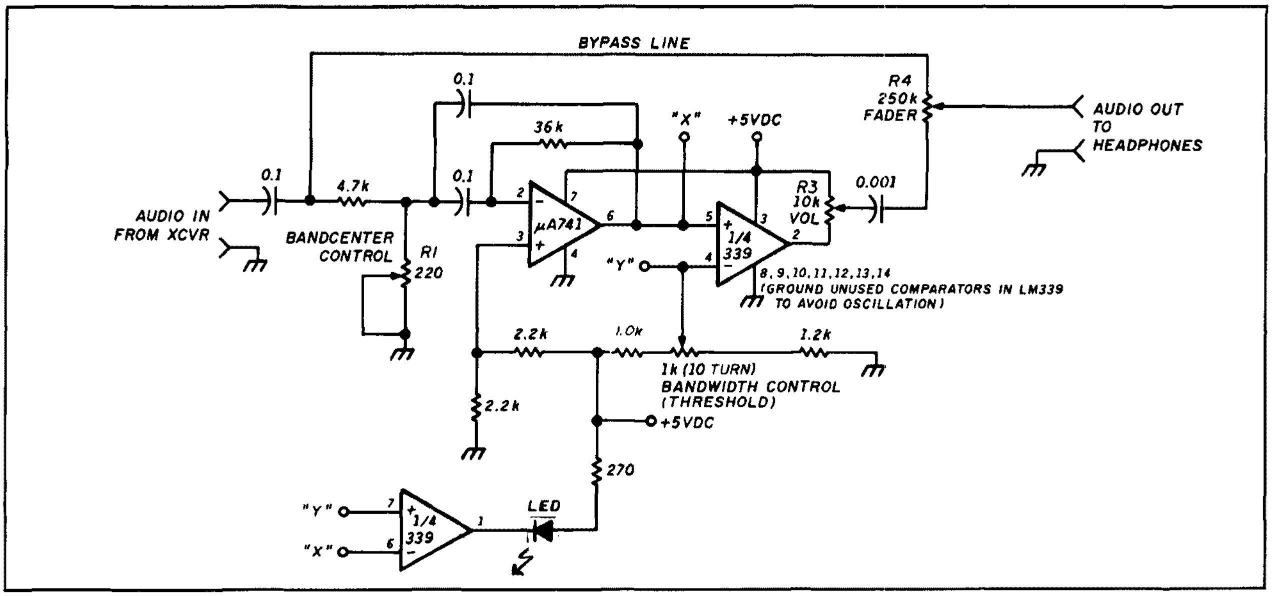
Figure 1 is the circuit diagram of my 75-Hz filter. The 741 integrated circuit used in the first stage is wired as a dual feedback bandpass amplifier with a Q of 10 and a gain of 4.[1] Normally, you'd need a multistage filter, with each stage limited to a Q of 4 or 5 to reduce ringing. However, the higher value of Q is possible here because only one stage is used. The result is a sharper, improved shape factor. R1 is a variable resistor for adjusting the center frequency of the filter to match the offset frequency of the receiver (typically in the 700 to 800-Hz range). If you intend to use the filter with only one receiver or transceiver, you can locate the frequency control internally to conserve panel space.

The second stage is a comparator whose output is low if the signal applied to pin 5 from the bandpass amplifier is lower than the value of the bias set on pin 4. The output voltage on pin 2 switches rapidly to the high state when the positive peak of the signal voltage exceeds the bias by a few millivolts. **Figure 2** illustrates the net bandpass characteristics of these two stages. The characteristics are those of an excellent CW filter. The flat top in the passband guards against ringing because pin 2 remains at a constant high level, even though the output of the bandpass amplifier varies above the bias setting. You get a shape factor of 1:1 because of the sharp change in output voltage when the signal goes above or drops below the bias set on pin 4 of the comparator. R2 varies the bias (threshold) level of the comparator, which in turn determines the bandwidth of the filter. Mount this control on the panel for operating convenience.

Another op amp is used for adjusting the comparator's bias. I selected an LM-339 because it requires a single polarity supply voltage and contains four op amps in its 14-pin IC package. This means that the second op amp (plus two spares) is already available. Its input is connected to the first op amp to sense the same audio signal and threshold setting (points X and Y in **Figure 1**). The output switches a light-emitting diode (LED) on and off electronically to indicate the threshold setting. The LED goes off when the background noise is biased out. It will then light as the code elements are received. Being able to see the received signal should be an aid when you're copying code.

The output of the comparator is connected to R3, which acts as a volume control for the filter. You adjust this control only once, so it can be located internally to save panel space. The wiper arm of R3 is coupled capacitively to one end of R4, which serves as a fader control. The other end of R4 is coupled back to the audio signal coming from the receiver. This arrangement lets you switch smoothly between unfiltered and filtered audio, or a blend of both. Locate this control on the front panel for easy access.

FIGURE 1



Schematic diagram of the complete 75-Hz bandwidth audio filter.

PARTS LIST

Recommended power supply:
Datel AC/DC regulated power adapter model PA-5/500.

Recommended cabinet:
Radio Shack no. 270-251.

All resistors 1/4-watt carbon composition or film types. All pots use linear tapers. All capacitors 25-volt disc ceramic types. Use shielded cables tor power supply and audio input leads to avoid RFI.

Design of bandpass filter[1]

Figure 3 shows how an operational amplifier can be wired as an effective bandpass filter. Solving three simple equations determines circuit values:

$$Q = (R3) \times \pi \times f_c \times C \quad or \quad R3 = \frac{Q}{\pi \times f_c \times C} \quad (1)$$

$$Gain = A = \frac{R3}{2 \times R1} \; or \; R1 = \frac{R3}{2 \times A} \quad (2)$$

$$R2 = \frac{Q}{(2 \times Q^2 - A) \times 2 \times \pi \times f_c \times C} \quad (3)$$

where: $C1 = C2 = C$
f_c = center frequency

$$Q = \frac{f_c}{3_{dB}\ BW} \qquad BW = bandwidth = \frac{f_c}{Q} \quad (4)$$

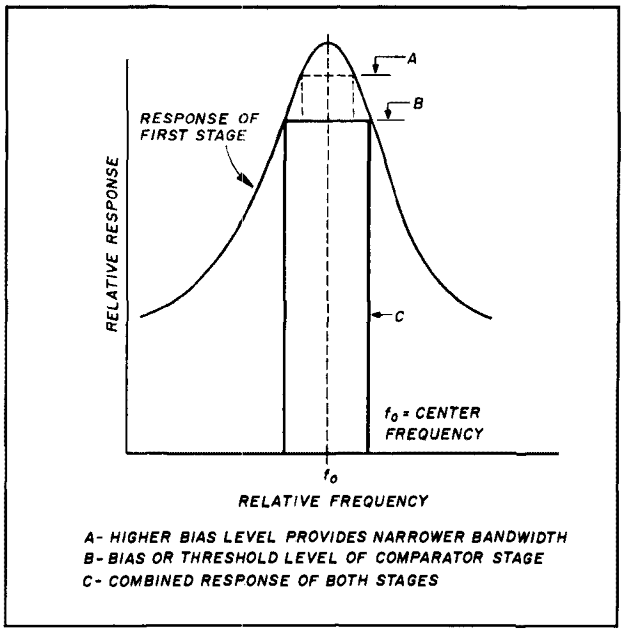
For: $f_c = 750\ Hz$
$A = 4$
$Q = 10$
Try: $C = 0.1 \mu F$

Then,

$$R3 = \frac{Q}{\pi \times f_c \times C} = \frac{10}{3.14 \times 750 \times 0.1 \times 10^{-6}} \quad (5)$$

$$= 42{,}441.3\ ohms$$

FIGURE 2



Illustrated response of the first two stages showing bandwidth adjustability and selectivity.

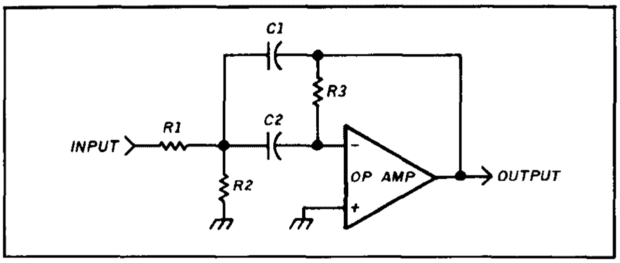
$$R1 = \frac{R3}{2 \times A} = \frac{42{,}441.3}{2 \times 4} = 5{,}305.2 \text{ ohms} \quad \textbf{(6)}$$

$$R2 = \frac{Q}{(2 \times Q^2 - A) \times 2 \times \pi \times f_c \times C} \quad \textbf{(7)}$$

$$= \frac{10}{(2 \times 10^2 - 4) \times 2 \times 3.14 \times 750 \times 0.1 \times 10^{-6}} = 108.3 \text{ ohms}$$

$$3_{dB}BW = \frac{f_c}{Q} = \frac{750}{10} = 75 \text{ Hz} \quad \textbf{(8)}$$

FIGURE 3



Basic schematic of an op amp (741) wired as an effective bandpass filter.

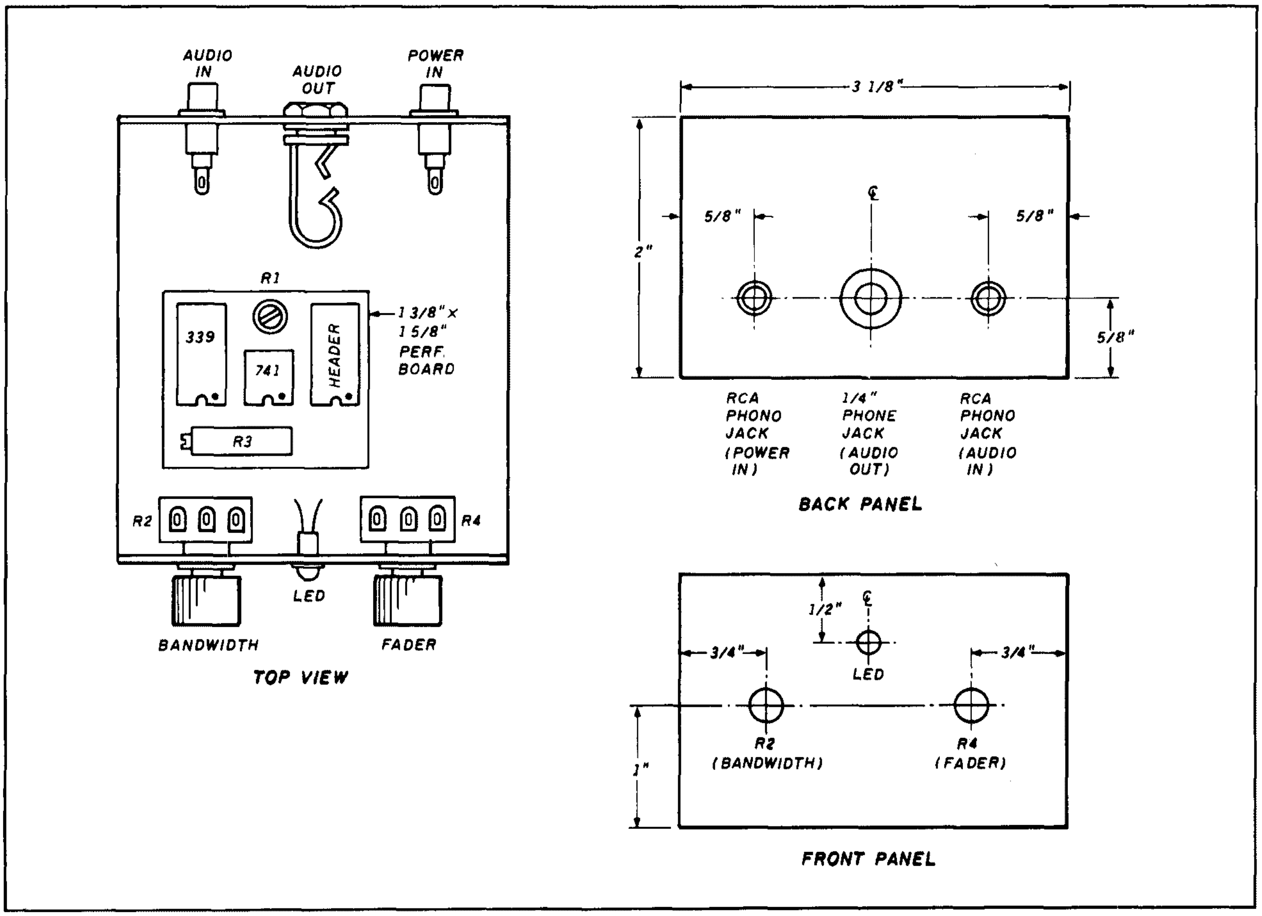
Because actual capacitor and resistor values tend to be on the high side of their tolerance ranges, select standard value resistors *below* the calculated values. For example, use 4.7 k for R1 and 36 k for R3. Variable resistor R2 can be any convenient value — like 200 ohms, or higher. A recalculation for 0.15-μF capacitors yields 3.3 k for R1, 100 ohms for R2, and 27 k for R3.

Construction notes

You can build the circuit in a variety of ways. There's nothing critical about the wiring. Point-to-point or pc board techniqes are proven methods; however, this project is ideally suited to a combination of both. All the fixed value resistors and capacitors can be soldered to a single 16-pin DIP header which plugs into a standard IC socket. Jumper wires from the socket to the panel-mounted components and active ICs complete the job.

Figure 4 shows the suggested arrangement of a 1-3/8 × 1-5/8 inch piece of perfboard, along with the parts layout on the front and rear panels. This layout is for use in a 3-1/4 inch wide × 2-inch high × 4-inch deep metal cabinet, like Radio Shack's catalog no. 270-251. The layout in **Figure 4** requires an external source of power. Take power from existing equipment or use a small wall-mounted power pack.

FIGURE 4



Mounting details and dimensions for the filter enclosure.

FIGURE 5

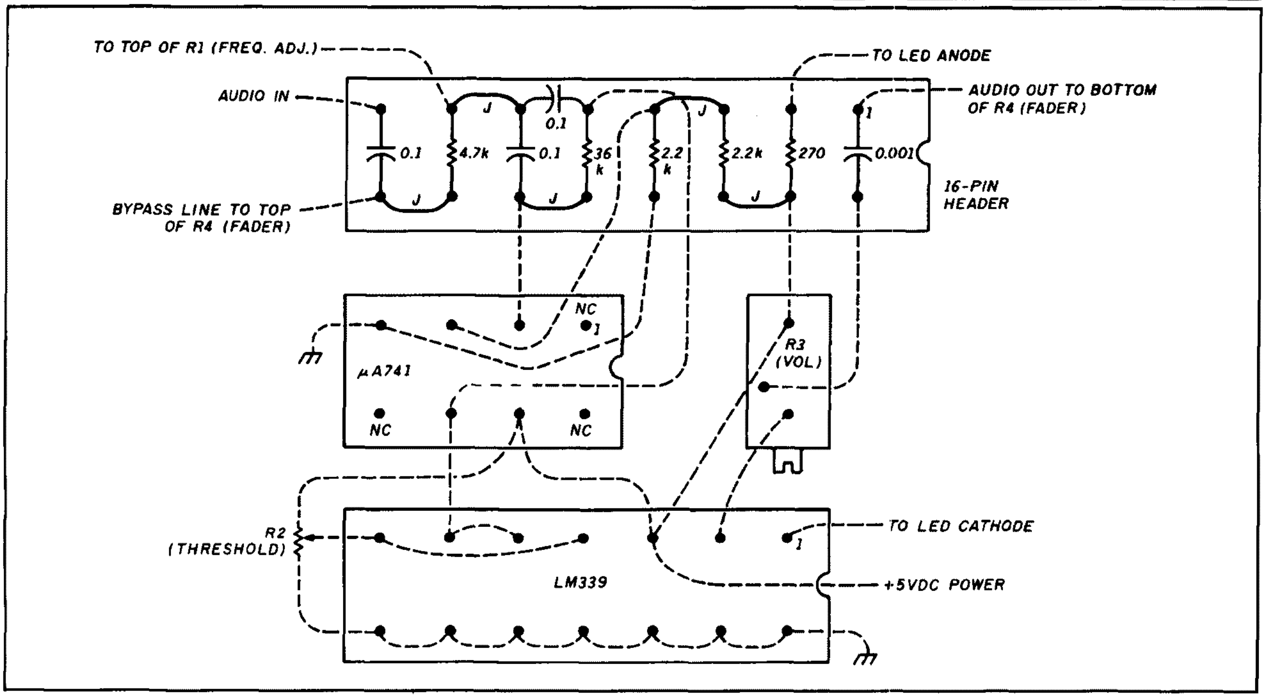


Diagram of parts placement and jumper connections for use with the DIP header type construction. Enlarged top view of header wiring and active ICs is shown. Heavy lines marked "J" are jumpers on header.

Figure 5 shows one arrangement for mounting parts on the IC header. The jumpers indicated by the heavy solid lines are also part of the header wiring. R2 can be any convenient value over 1 k. If you use 5 k or more, the 1-k and 1.2-k resistors shown connected to R2 in **Figure 1** can be eliminated.

Adjustment and operation

Adjustment of the four variable resistors is simple and straightforward. R1's setting determines the center frequency of the bandpass amplifier. If you plan to use the filter with a single receiver or several pieces of equipment with the same offset frequency (CW beat note), R1 is a "set and forget" control. Tuck it away inside the cabinet to conserve panel space. The same applies to R3, which is used to adjust the volume of the filter's output to match that of the receiver. R2 and R4, however, should always be located on the front panel. R2 determines the switching threshold of the comparator stage, and its setting is influenced somewhat by the strength of the received audio. **Figure 2** shows how the setting of the threshold level in relation to the response curve of the bandpass amplifier determines overall bandwidth. Note that the filter's selectivity narrows as the threshold is set closer to the peak of the curve. I recommend that you use a multiturn potentiometer, although a conventional control with about 330 degrees of rotation is satisfactory.

R4 acts as a fader control, but you can replace it with an SPDT switch. Unfiltered audio appears at the output of R4 when the wiper arm is pointing toward the bypass line. This position is handy for general tuning purposes. It's advantageous to use a variable control, rather than the switch arrangement, because you can move the wiper away from the full bypass position in smaller increments. In this mode of operation the narrower bandwidth still permits casual receiver tuning, but adds sufficient resistance to the bypass line to force some audio to enter the filter and activate the LED. Moving the wiper to the other end of R4 provides fully filtered audio at the output jack. In this position, the quality of the audio note changes from the original smooth sounding sine wave to a highly clipped note reminiscent of the CW limiters in old vacuum tube receivers. The resulting sound approaches that of a square wave and, in my opinion, produces a CW note that's easier to copy.

Adjust R1 using a steady carrier picked up by the receiver. You can use a crystal calibrator to supply the carrier. Tune the receiver for maximum deflection of the S-meter and leave it undisturbed while you make the rest of the adjustments. With the receiver audio level set no higher than needed for detection by the filter, and both R1 and R2 set fully counterclockwise (maximum bandwidth), the LED should be on. Carefully turn R2 clockwise until the LED goes off. Adjust R1 until the LED comes on again, and note the knob position. Continue turning R1 *in the same direction* until the LED goes off; note the knob position again. Return R1 to the setting midway between the transition points and turn R2 clockwise slightly to extinguish the LED. Reduce the filter bandwidth progressively using R2 and adjust frequency control R1, while maintaining minimum necessary excitation from the receiver. Eventually, you'll reach an adjustment of R1 which matches the filter's center frequency exactly to the offset frequency of the receiver. *Once optimized, R1's setting shouldn't be disturbed* during normal filter operation.

This filter is very easy to use. Just plug it into your headphone jack and turn up the receiver's volume control until the LED blinks in unison with the incoming code elements. Keep the fader control aimed toward the bypass line for maximum bandwidth. As I mentioned earlier, this position is best for general tuning purposes — including listening to phone QSOs. Moving the fader slightly away from the broadest position reduces the bandwidth enough to filter out background noise and cut phone QRM. Turning the fader control to the sharpest position lets you copy CW signals in the noise, or through heavy interference. The filter's sharp skirts will amaze you. A signal tuned to the edge of the passband simply disappears.

Acknowledgments

Part of this filter's circuitry was adapted from an article written by David Jagerman, KC2FR.[2] I'd also like to thank Jack Thompson, W2OPE for his valuable assistance in building the first prototype. ■

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1. Don E. Hildreth, W6NRW, "Graphic Filter Design," Figure 2, *Ham Radio*, April 1984, page 41.
2. David Jagerman, KC2FR, "The KC2FR QRM Fighter, *QST*, July 1982, page 28.

TV TO 2-METER ANTENNA CONVERSION

By Karen D. McIntyre, N4FQO, 3711 Gayle Avenue, Omaha, Nebraska 68123

As a very new ham, licensed for just a month, I found the idea of building a 2-meter antenna rather intimidating. However, my XYM (Ex-Young-Man) said that if I wanted one, it was up to me to build it. Thus challenged, I set to work.

I had originally planned a simple three-element beam, but because I find it impossible to keep anything simple, the antenna grew to five elements before I was out of the planning stage. It's an amazingly simple antenna to build — even for a rookie. All you need is three free hours, a scavenged TV antenna, some stainless steel hardware (nuts, bolts, flat and lockwashers, and two solder lugs), 7 feet of RG-58/U coax, a PL-259 coax connector, and a few strips of flat metal or wire (for the beta match).

This is also an inexpensive antenna to build, though the actual cost will depend on your scavenging ability. In my case, it only cost me $3.00 for some hardware I couldn't find in my husband's junkboxes. I garnered my TV antenna by climbing up on our shed and unbolting an unused antenna abandoned by the previous owners. My husband realized he had created a monster when I dragged it into the dining room to begin work on my creation! The fruits of my labor are shown in Photo A.

Selecting the antenna

You can use almost any TV antenna of the Yagi or log-periodic type, provided that it's in fair-to-good condition. Insulate the driven element from the boom. Don't use a folded dipole without a matching transformer.

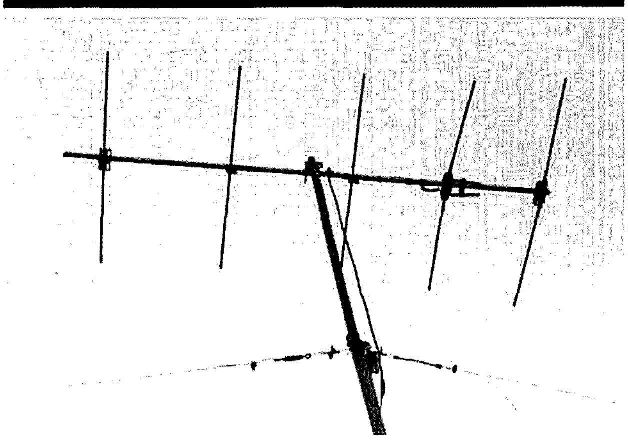
Make sure the TV antenna you choose has at least 71″ of boom. Any excess length may be either left alone or cut off; it makes no difference in the performance of your antenna. You'll need to salvage five elements of about 40″ each for the elements of your 2-meter antenna.

Disassembling the antenna

Once you've rounded up the TV antenna, the next step is to take it apart. Before you start taking the elements off the beam, look at Figure 1. It shows you the finished lengths of the elements and their spacing on the antenna. If possible, leave the first element (the longest one) of your TV antenna in place. This becomes the reflector of your antenna. Simply cut off the ends of the element to obtain the desired length (39″ tip to tip). It doesn't matter if the element is attached to the boom with a metal bracket; all the elements except the driven element are grounded to the boom, anyway.

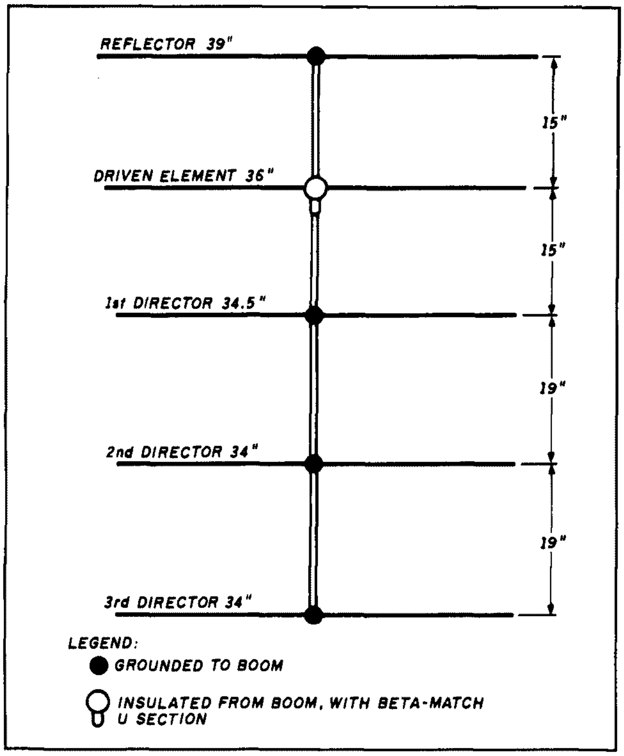
Now remove the remaining elements from the TV antenna. Save all the brackets and hardware that you can and reuse it where possible during reassembly. You may have to drill out rivets to disassemble the TV antenna. Be careful not to damage the brackets when you do this, as they are hard to find and quite expensive.

PHOTO A



Completed 2-meter antenna. (Dimensions shown are for 146.50 MHz.)

FIGURE 1



Driven element details.

Assembling the antenna

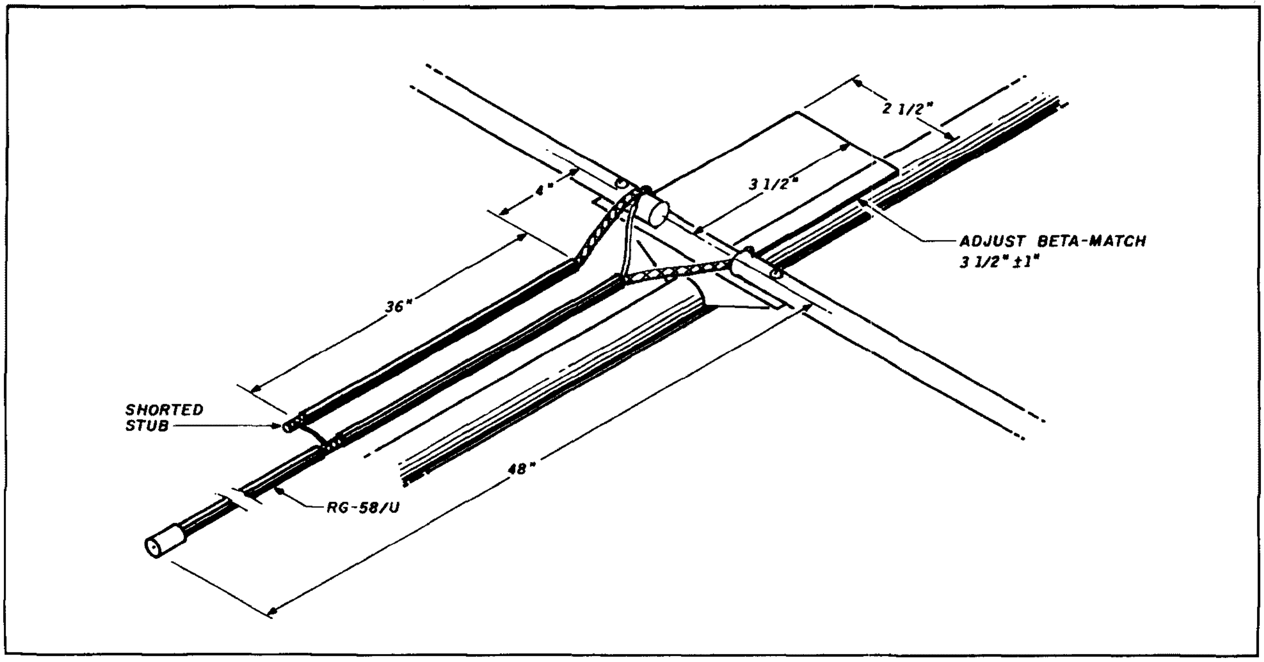
You now have a pile of salvaged elements and hardware, and a boom with one element attached. So far, so good. You're ready to put everything back together. If possible, use stainless steel hardware when reassembling your antenna; it's much more wear and corrosion resistant.

The next element, the driven element, must be insulated from the boom as it is "hot." The TV antenna I used had plastic brackets attaching three of the elements, so I simply used one of those. If your antenna doesn't have any plastic brackets, rummage in the junkbox or make one out of a piece of plastic or Lucite™. Measure 15" along the boom from the center of the reflector and drill a hole through the boom. Attach the plastic bracket to the boom, and any two element pieces to the bracket, with long stainless steel bolts, nuts, and lockwashers. Include two no. 10 stainless steel solder lugs when you affix the driven element, and tighten this piece just a bit for now. The solder lugs are used for connecting the coax line later. The beta match will also be attached at this point. Now cut this element to measure 36" total, tip to tip.

The last three elements — the first, second, and third directors — are virtually identical. Refer to **Figure** 1 for spacing and element lengths. Simply drill a hole through the boom and mount a piece of element with stainless steel hardware and scavenged metal brackets. Then trim the elements to the correct lengths.

I think a few words are called for here regarding the accuracy of your measurements. Relax! Precision to the second and third decimal points isn't necessary, though the formulas in most antenna handbooks would seem to indicate

FIGURE 2



Two-meter antenna made from salvaged TV antenna, assembled and ready to roll.

otherwise. Stay as close as you can to the recommended lengths, but remember that a slight variance of 1/8″ or so here or there isn't going to ruin your antenna. After I had cut and attached all the elements, I discovered that my old and frayed tape measure gave a slightly different reading each time I measured the same piece of metal. My husband, responding to the wails of distress emanating from the dining room, assured this panic-stricken rookie that everything would be okay — but I didn't really believe him at this point.

Matching network

The last order of business is the construction of the beta match and coaxial balun shown in **Figure 2**. The beta match is nothing more than a U-shaped conductor bridging the gap across the driven element. I used some flat pieces of metal 1/2″ wide by 1/8″ thick scavenged from the TV antenna and secured in a U shape 2-1/2″ wide by 3-1/2″ long. However you make the beta match, be sure that you can adjust its length (from 2-1/2″ to 4-1/2″ overall) for the best VSWR.

Now construct the coaxial balun as shown in **Figure 2**. Attach a PL-259 connector to one end of a 4′ length of RG-58/U coax. At the other end of this 4′ length of coax, separate about 4″ of shield and center conductor. Measure back 36″ from the point where the shield and center conductor divide, and expose about an inch of the shield by removing the outer insulation all the way around the coax. Cut another 36″ piece of coax; solder one end of this to the exposed shield of the feedline and the other end to the center conductor (yes, center conductor!) of the feedline. Cover all solder joints, braid, and the center conductor with tape or heatshrink tubing to keep moisture out and to keep the sun from cracking the exposed center insulation. Attach this completed feedline to the solder lugs on the driven element, tighten the hardware, and tape the cable securely to the boom. I know this sounds like one big short circuit, but it works!

Adjusting the antenna

Now it's time to test your craftsmanship. Attach the feedline from your station to the PL-259 connector on the antenna feedline and check the VSWR. Through some blessed combination of skill and beginner's luck, my antenna came up at 1.2:1 VSWR at 146.97 MHz with very distinct front-to-back ratio the first time I tried it. (My XYM wanted to tinker with it, but I threatened him with bodily harm!) If you're not as fortunate, use the following steps to adjust the antenna:

- Adjust the driven element length for the lowest VSWR at your preferred operating frequency.
- Adjust the length of the U-shaped beta match until the VSWR at this frequency is minimum. You should have no trouble bringing it below 1.5:1.
- If you are a purist, repeat the two steps above, as there may be some iteration.

Summary

Mount your antenna any way you prefer. Now fire up the rig and have some fun! This antenna gives excellent performance, is lightweight, rugged, and has survived three Navy moves. It was a great confidence-building project for a rookie ham.

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MAKING PRINTED CIRCUIT BOARDS

Don't give in to perfboard just yet!

By Dave Mascaro, WA3JUF, RD 1 Box 467, Ottsville, Pennsylvania 18942

There are many ways to produce your own printed circuit boards. Small boards for DC switching circuits are easy to make because neither the size nor the shape of the copper foil traces is critical. Boards containing many digital ICs are considerably harder to lay out and etch. There are also some double-sided digital board designs with plated-through holes. These types of boards aren't reproducible and you're better off buying them from the source.

Many articles have been published on the use of stencils and rubyliths in conjunction with sensitized pc boards and light tables. These techniques are good for the more complicated boards or for making multiple boards. The process involves making a positive or negative artwork, exposing and developing the light sensitive board, and then etching the board. Most homebrewers don't have access to light tables, developing solutions, and darkroom facilities. I use light sensitive methods for many of my pc boards — mostly for digital IC boards or multiple boards of the same design. What follows are some pc board techniques you can use instead of these facilities.

There are ways to produce RF microstrip and small DC pc boards in minutes without light tables, some without the etching process. Basically, there are four methods. Each method has its own variations, depending on the board material.

- X-ACTO™ knife method
- X-ACTO etch method
- Drafting tape method
- Dremmel method

These methods involve the use of an X-ACTO knife and other procedures that require safety glasses or goggles. The tip of an X-ACTO blade can snap off when used to cut copper foil. I purposely snap off a very small portion of each new blade before starting. This way I'm sure that the small blade tip will end up in the trash can and not on my workbench.

X-ACTO knife method

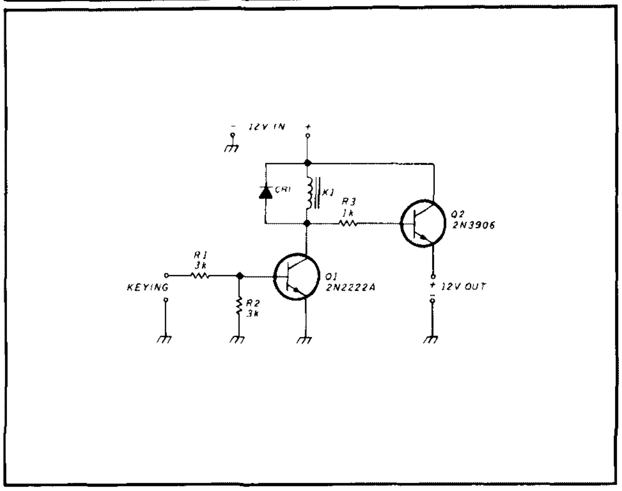
This type of pc board construction can be used on standard G-10 fiber glass board or Teflon™ dielectric boards. It involves cutting away strips of copper, leaving isolated pads to which components are soldered. This is a non-etch process for making pc boards quickly. A working project board can be completed in a quarter of the time it takes to make just the pc board using light sensitive methods. A pencil soldering iron with a pointed tip is required.

Figure 1 shows a small DC switching circuit to be made on G-10 board. Parts placement isn't critical on a simple non-RF board, and the parts layout can follow the schematic directly. Islands of copper are needed at all junctions of the individual components. Ground connections are common using the main copper foil portion of the board.

Figure 2 shows the pc board layout for the circuit in **Figure 1**. Using a pencil, draw small rectangular pads on the copper foil for each component connection. Draw a second outline slightly larger than the first. This small strip of copper around each pad will be removed later. You can draw symbols for each component on the board to make layout easier and ensure correct spacing for leads.

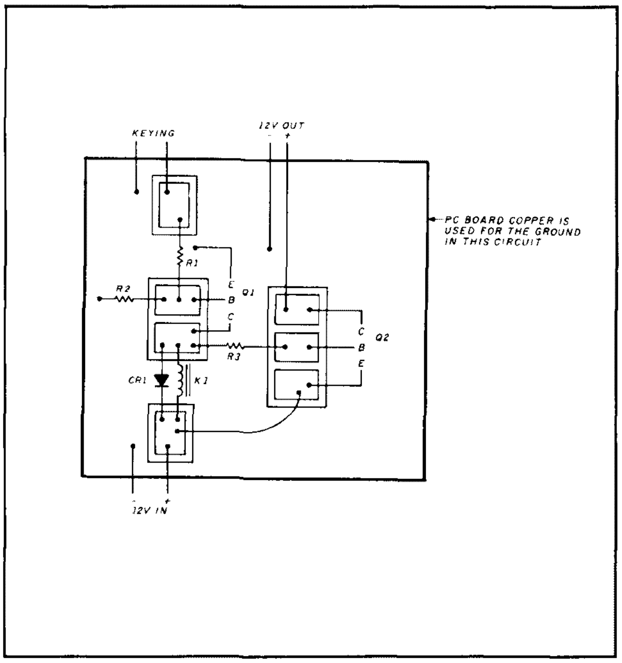
After drawing all the pads, mark and drill component mounting holes. Now, while wearing safety glasses, use the X-ACTO knife to cut along the pencil lines. Use a straight edge if you wish. Heat up the small strips of copper by dragging the tip of your soldering iron back and forth along the strip. The copper foil will pull away from the fiber glass board easily. You can remove very narrow strips with the soldering iron tip. Make the strips just wide enough to isolate the

FIGURE 1



Simple DC switching circuit.

FIGURE 2



Board layout showing pads made by removing copper strips around each square using X-ACTO knife methods. A "score knife" could also be used to cut away the copper, leaving the isolated pads.

pads. Wider foil strips can be tinned with solder first, then removed using the X-ACTO knife along with the soldering tip. After all the pads are isolated from ground, use steel wool to smooth the board and remove any oxidation that would hinder soldering the components in place.

You can remove copper to make mounting pads in another way — with a "score" knife. Roofing and siding people use this type of knife to cut aluminum siding. The blade makes a "V" cut in the aluminum, making it easy to bend and break. The blade will also cut a V-shaped groove in the fiber glass pc board. Using this knife is sometimes much easier than using an X-ACTO knife because you need to make only one cut. Dragging the knife along a straight edge allows you to remove long strips of copper for long rectangular pads. With this knife and a straight edge, you can cut pads with 0.1-inch spacing to accomodate ICs, computer-type connectors, or header plugs. I made a board with seven ICs, two LED readouts, and three header plugs using the score knife. Making pads with IC spacing using this knife, in conjunction with some wire wrapping, is a simple way to make an otherwise complicated pc board. After cutting the board to the proper layout, drill component and IC holes and use steel wool to smooth out the knife cuts.

You can use a double-sided pc board (copper foil on both sides) on larger boards with many grounding points. On double-sided boards, the non-etched or non-patterned side of the board is referred to as the ground plane side. All grounds are connected to the ground plane side of the board by holes drilled for each ground lead. Isolate the non-ground leads on the ground plane side of the board by clearing the holes with an oversized drill bit. Drill and clear all non-ground holes first to eliminate the confusion of which holes get cleared.

The X-ACTO knife method can also be used on Teflon pc boards for very basic microstrip designs. **Figure 3** shows a simple RF amplifier. A pc board for the amplifier with a 50-ohm microstrip is shown in **Figure 4**. This board can be produced using the X-ACTO knife method because all lines are straight. Basically, you need remove only two strips of copper. A working amplifier of this type can be completed in 15 minutes using X-ACTO etch.

The copper on a Teflon pc board won't peel when heated with a soldering iron as it does on G-10 board. Peel off the copper using the knife and good needle nose pliers. After drawing the design on the board with a pencil, cut along the lines with the knife. Cut just deep enough to go through the copper foil. Now use the X-ACTO knife and needle nose pliers to pull up the foil strips. After doing a few boards this way, you'll learn to remove the foil without cutting into the Teflon dielectric. Use other techniques for more complicated boards.

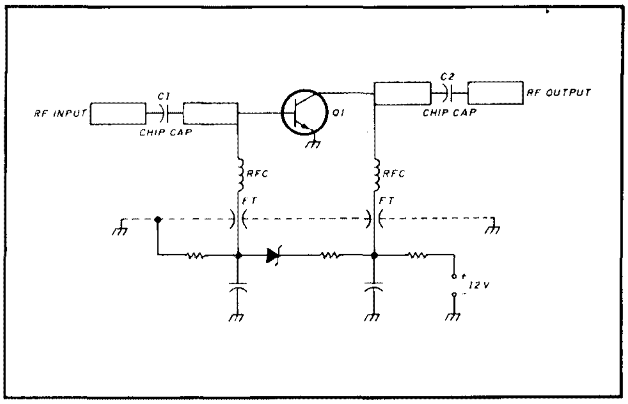
An X-ACTO knife is all you need to make microstrip-type boards on Epsilam 10-type board material (dielectric constant of 10). You can cut the foil with the knife and peel it off with a small pair of needle nose pliers. It's easy to remove the copper foil from this type of board material.

X-ACTO etch method

This process is used for making microstrip-type pc boards. The technique takes little time, doesn't require photo-type artwork, and eliminates the exposure/developing processes. With X-ACTO etch, you prepare your pc board for etching by cutting "clear tape resist" to the desired pattern with an X-ACTO knife. You can purchase clear tape on a 4-inch wide roll at most stationery stores. Although light sensitive methods would normally be used for producing multiple boards, I often use X-ACTO etch because it's faster. It takes less time to X-ACTO etch two or three boards than it does to use light sensitive methods. Making the rubylith artwork for light sensitive methods can be very time consuming.

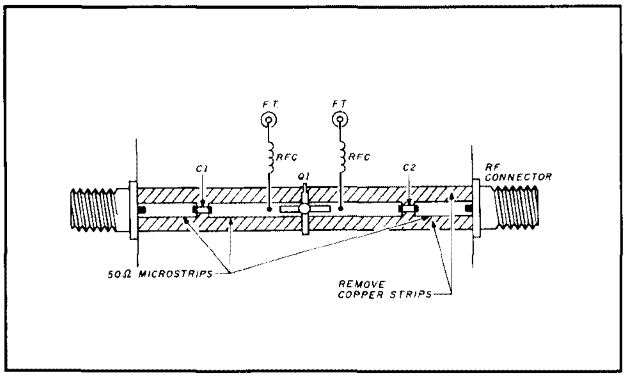
Select a pc board of the proper size. You must use double-sided board for all microstrip designs. Draw the desired pattern on the board with a pencil and straight edge. You can use a dial caliper to accurately lay out a

FIGURE 3



RF amplifier using 50-ohm microstrip.

FIGURE 4



Layout of RF amplifier showing two strips of copper to be removed with an X-ACTO knife to produce a microstrip pc board. Cut away copper to mount DC blocking chip caps C1 and C2.

design from full scale artwork in a magazine. Most microstrip designs have trimmer capacitors for optimization, anyway.

Cover both sides of the board with clear tape. Cut the tape along the penciled lines, using a straight edge as needed. Remove the tape where the copper is to be etched. The remaining tape will serve as the resist for etching the board. Press the resist tape firmly to the board to reduce the "under etching" effects of the ferric chloride. You can add any additional pads to the board later with pieces of tape. Etch the board and remove the resist tape. Clean the board with steel wool and trim to size if necessary.

Another variation of X-ACTO etch involves using a photocopy of the full-scale artwork as a pattern from which to cut the tape resist. Attach the board to the reverse side of the photocopy with tape. Cut the resist tape with your knife, following the artwork copy. You may find this technique difficult to use on some designs because, depending on the number of cuts, the artwork copy will fall apart before the last cuts are made. The last few cuts can usually be made by taking measurements with a dial caliper. After the tape pattern is cut, continue by using the procedures mentioned earlier.

Drafting tape method

You can make many DC switching and small IC projects on single-sided pc boards with this technique. Drafting supplies for making light table artworks are used as the actual resist for etching a board. Drafting tapes are available in many different widths. "Donuts" and other pad shapes are used to mark drill holes. Transistor and IC patterns are also available. Use clear tape to make your own special patterns. Then use these patterns in conjunction with drafting tapes to create circuit resist patterns on the copper side of the board.

Press the patterns and tape lines to the board firmly. Some under etching will occur, but this isn't a problem on DC-type boards. Making the lines a little wider initially will help. Now etch the board, remove the resist tape, and clean the board wih steel wool.

Dremmel method

A dremmel tool or small drill press can be used to cut away the copper foil to produce small pads of copper on G-10 fiber glass board. Instead of using the X-ACTO knife and soldering iron, use a small dremmel bit to cut isolating pads on the board. You can do this freehand on boards that don't require a neat appearance.

I use small dental drills in a drill press to cut some of my boards. The worktable is raised up close to the drill bit and set so the bit cuts only through the copper foil. Mark the copper side of the board with a pencil. Move the pc board along the worktable, following the penciled lines. A straight edge clamped to the worktable acts as a guide, letting you make straight cuts on the pc board.

Etching boards

You can etch boards in several ways. The easiest method is to use a spray etching tank. If you don't have access to a tank, the method that follows can be used for basement or garage etching. Conventional methods involving plastic trays with agitating motors and heat lamps work too.

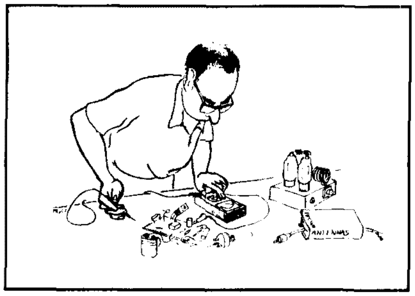
Place the board to be etched in a Tupperware® or other sealed plastic container with sufficient etching solution. Ferric chloride works best at higher temperatures. The etching solution should be less than half the volume of the container to allow for air expansion. Put the plastic container in a bucket of hot water and agitate until the board is etched.

Another etching container that works well is the ceramic "crock pot." I got the idea of using the crock pot from Phil, WA3NUF. Pour about 1 inch of ferric chloride into the pot. Put your board in the pot and replace the cover. Set on the low heat position. Your board will be done in 30 minutes. You can etch several small boards at the same time, providing they aren't stacked on top of each other. The ferric chloride fumes stay in the pot; this is an advantage over using a plastic tray for etching. This type of etching can be used for all resist methods. Even in the high heat position, the clear tape resist works just fine.

Conclusion

I hope these methods will encourage you to build projects requiring a pc board. A double-sided board with plated-through holes is nice, but isn't always necessary. Next time you think about a perfboard project, try it on a pc board using the X-ACTO knife method. I think you'll find it's easier

Ham Radio Techniques



Bill Orr, W6SAI

ANTENNA GAIN

Antenna gain is nice. Those who have it are louder on the air than those who don't. It's as simple as that. Antenna gain costs money and takes up space. Antenna gain, much like "music power" in a stereo amplifier, is a concept that brought joy to the believers in the hard sell technique. Antenna gain advertisements touted bigger and better gain figures that stretched the imagination and glazed the eyes of the innocent.

A decade or so ago, the "gain wars" between antenna manufacturers grew so intense that the gain figures given in antenna advertisements were no longer accepted in some Amateur magazines. Things have calmed down now, and sophisticated hams have a pretty good idea of the gain available from various antennas. They also know the expressions that define the gain.

Expressing antenna gain reference in dBi and dBd

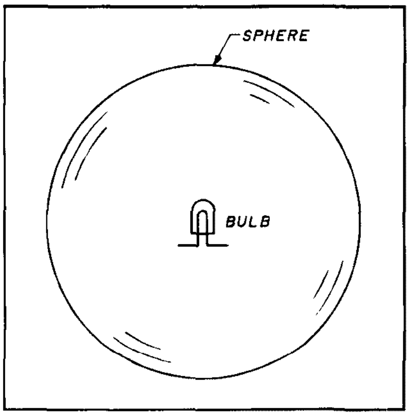
For the record, accepted expressions of antenna gain reference either an isotropic radiator or a comparison dipole. The isotropic radiator is a theoretical antenna that radiates equally well in all directions. This concept can be illustrated by a tiny light bulb placed at the center of a large sphere. The bulb illuminates the interior of the sphere equally at all points (**Figure 1**). Antenna gain in decibels referenced to an isotropic radiator is expressed as dBi.

A real-life antenna has some degree of directivity and, if placed in the center of a large sphere, illuminates certain portions better than others. That is, the antenna radiates energy better in certain directions. The simple dipole has maximum radiation at right angles to the wire and minimum radiation off the ends. Antenna gain in decibels referenced to a dipole is expressed as dBd.

If the same amount of energy is radiated by the dipole as by the isotropic radiator, the energy in the direction of maximum radiation of the dipole is about 2.14 dB greater than that of the isotropic antenna. So it can be said that a dipole has a power gain of 2.14 dB in the direction of the main lobes over an isotropic radiator.

All of this supposes that the isotropic radiator and dipole are in free space. What happens when an antenna is placed in proximity to the earth?

FIGURE 1



An isotropic radiator is simulated by a tiny light bulb placed at the center of a large sphere. The bulb illuminates the interior of the sphere equally at all points.

The effect of the earth on antenna gain

The best way to visualize the antenna near earth is to cut the sphere of measurement in half horizontally and substitute a conducting metal plate for the missing portion (**Figure 2**). Now, only half of the original sphere is illuminated. The other half is represented by the ground plate. Assume a horizontal half-wave dipole is placed at the center point of the sphere. The direct and reflected waves are in phase, or additive. Because the same amount of power is radiated in half the volume, the field is *doubled* — or increased by 3 dB. In addition, if the ground plane is smooth and a perfect conductor, ground reflection adds another 3 dB. Thus, a dipole in a "near earth" situation can have up to 6-dB gain over the antenna in free space.[1,2]

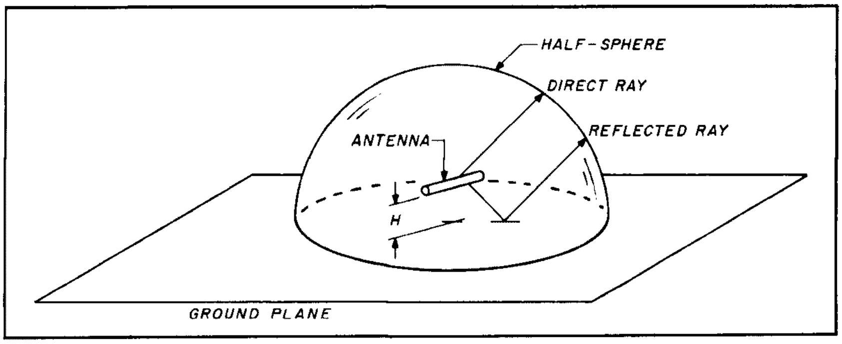
Note that I said "up to." Maximum ground reflection gain occurs in line with the direction of maximum radiation. Cancellation of gain occurs at other angles above the horizon. Limiting cases are either 6 dB greater than the free space value, or zero. The result is the well-known radiation pattern shown in **Figure 3**.

The reflection gain figure can be less than the theoretical value if ground conductivity is poor. Ground conductivity studies have been made since the early days of broadcasting, and the values in the United States are well known.[3] Conductivity is expressed in siemens (or millisiemens) per meter. The dielectric constant is also given (see **Table 1** in **Reference 3**).

There's a computer program* available that lets you examine reflection gain as a function of ground conductivity (**Figure 4**). In an area of high ground conductivity, the reflection gain for a dipole is better than 6 dB over the free space value. As ground conductivity and dielectric constant decrease, the reflection gain decreases. In the

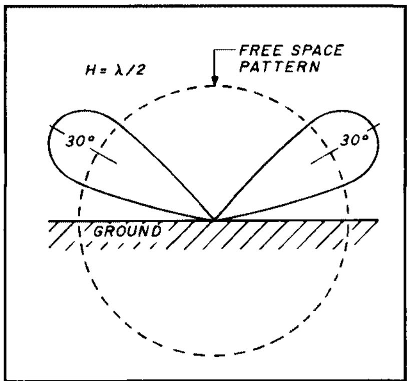
*MN Antenna Analysis, Brian Beezley, K6STI, 507-1/2 Taylor, Vista, California 92084.

FIGURE 2



The effect of ground is simulated by placing the antenna over ground plane. Because the same amount of power is radiated in half the sphere volume, the radiated field is doubled. Ground reflection depends upon ground conductivity and can increase the field by up to 3 dB in line with maximum radiation.

FIGURE 3



Ground modifies the free space pattern. This shows a half-wave dipole 1/2 wavelength above ground. Six dB gain is added when antenna is above perfect ground, as compared with free space.

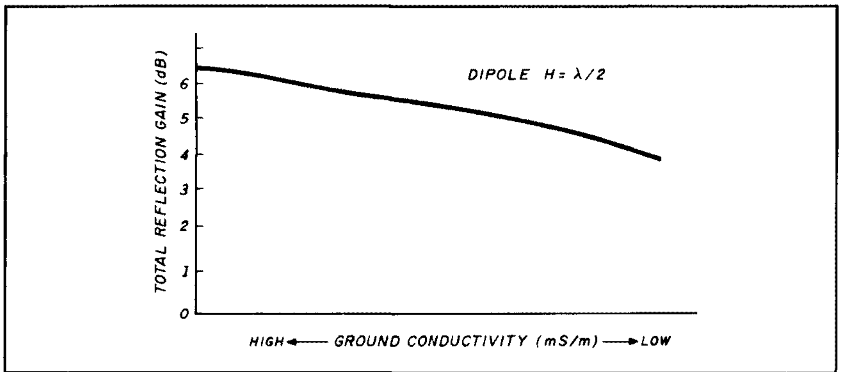
example shown, the dipole loses nearly 3 dB in reflection gain when going from an area of high ground reflection to an area of poor reflection.

Enter the computer

The conventional dBi and dBd gain figures are well known, but the picture has been clouded recently by computer-derived antenna programs that automatically add up to 6-dB ground reflection gain to the results — depending on the value of ground conductivity you enter into the program. This is producing eye-popping gain figures for even rather simple antennas. A casual observer might jump to the tantalizing conclusion that a dipole is indeed a "gain" antenna. After all, it has up to 6 dB of ground reflection gain, doesn't it?

Thus, there are four valid ways of referencing antenna gain: (1) in free space, referenced to an isotropic radiator; (2) in free space, referenced to a dipole; (3) in proximity to the earth, referenced to an isotropic radiator. (4) in proximity to the earth, referenced to a dipole.

FIGURE 4



As ground conductivity decreases, ground reflection gain also decreases.

Comparing apples and oranges

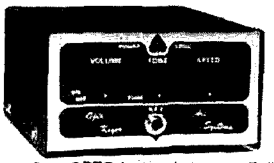
Everything seems simple enough, up to this point. But the possibility exists that some antenna manufacturers will claim inflated gain figures. The opportunity to tout reflection gain can be irresistible. All's well if everyone uses the same ruler for gain comparison. I've noticed, however, that one antenna manufacturer has latched onto the ground reflection concept and is boasting antenna gain figures based on earth proximity, referenced to an isotropic radiator. His data sheet doesn't mention any definition of the gain reference.

More and more computer-designed antennas are being publicized, and it's common to see antenna gain figures based upon earth reflection.[4] Once the concept of reflection gain and ground conductivity are well known, and antenna gain figures are properly referenced, the computer-derived gain figures fall into their proper perspective.

Halyard replacement

I have a bi-square loop antenna[5] supported by a halyard passing over a pulley that hangs from a yardarm near the top of my tower. The rope was quite frayed and I wanted to replace it before it broke. I knew that if it did break, it would run through the pulley, and I would have to climb the tower and hang by my heels to get a replacement rope through. Because fear sets in if I go more than 10 feet above the ground, this was an unpleasant thought!

How was I going to get a replacement rope through the pulley from ground level? That was the puzzle. I couldn't splice the new rope to the old and pull it through the eye of the pulley. There was no room for that! I had forgotten long ago the tricky knots that earned me a Scout merit badge. Finally, I had a brainstorm! I could use heatshrink tubing to splice the rope ends together! I placed the end of the old rope in one end of a foot-long section of heatshrink tubing; the new rope went in the other end. I used safety matches to shrink the tubing. The two



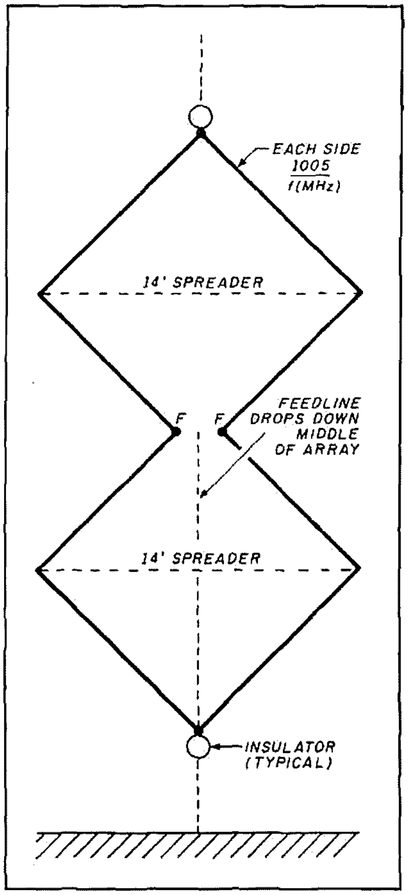
ropes were firmly connected and the splice slid easily through the pulley! Q.E.D.

The dual quad loop antenna

A single, vertical quad loop makes an effective antenna. It has the radiation pattern of a dipole and provides an additional gain of approximately 1.2 dB. The quad loop has a very broad band response and a feedpoint impedance of about 120 ohms. Place two of these loops in phase and feed them at the common point, and you have Jeff O'Connell's, K4BLT, dual quad loop antenna.

Jeff's antenna is cut for the 12-meter band. It's suspended from a branch of a pine tree (Figure 5). Two oak spreaders 1-1/4 inch square and 14 feet long form the diamonds. The pattern is bidirectional and the gain is estimated to be about 3.5 dBd. Polarization is horizontal.

FIGURE 5



Antenna is fed at F-F with coax line. Line is wound into four-turn RF choke at feedpoint.

The antenna is fed at the center with a coax line. The feedpoint impedance is very close to 50 ohms. The line is wound into an RF choke at the feedpoint. The choke consists of four turns of coax 5 inches in diameter. This helps keep RF off the outer shield. Bring the line down the middle of the array, as shown.

The K4COF dragonfly effect

Phil Elrod, K4COF, reports a puzzling effect he noticed on a dipole antenna when he was stationed in Japan. He says that he saw dragonflies sitting on the antenna, all facing in the same direction. Each time the transmitter was keyed, the dragonflies would instantly leap off the antenna and hover in the air. When the transmitter was unkeyed, the insects would return to the wire! He said the dipole was fed directly with coax; no balun was used. The hopping insects were mainly on the dipole half connected to the center conductor of the coax!

It occurs to me that dragonflies might be good, cheap indicators of antenna balance. They might even be more effective than an SWR meter! However, there aren't many dragonflies in northern California. How about hummingbirds or killer bees? Do you suppose Phil is pulling my leg?

The "Dead Band" Quiz

K4COF also supplies this month's quiz. (He doesn't supply the answer because he doesn't know it!) The question is: At 4:00 and 8:00 o'clock the clock's minute and hour hands form a 120-degree angle. Is there a time at which the hour, minute, and second hands form three simultaneous 120-degree angles? If you know the solution, and can prove it, drop me a note at Box 7508, Menlo Park, California 94025. Phil will get his answer when he reads this column.

Coming next month: results of the "snowplow" quiz! **hr**

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COPPER PIPE TRANSMISSION LINES

"Plumber's delight" antenna feed system

By Ernie Franke, WA2EWT, 10484 138th Street N., Largo, Florida 34644 and Judd Sheets, WA4BUS, 6926 10th Avenue N., St. Petersburg, Florida 33710

Copper pipe offers a mysterious attraction to the VHF/UHF enthusiast. When he comes across a large diameter piece of tubing in a scrap yard, he has a compelling desire to store it away in anticipation of building a cavity.

The tables in this article show the impedances obtainable when using copper pipe from a plumbing store. You may find them helpful in determining how to use these pipes to form coaxial TEM structures. Copper pipe has been used for coaxial transmission lines,[1] antenna power dividers,[2] coaxial filters,[3] cavity duplexers,[4,5] and baluns.[6] By considering the best impedance for achieving the lowest loss, you can design these structures for the highest Q and minimum loss.

The design of a transmission line or a coaxial cavity is set by the maximum allowable inside diameter of the outer conductor (D), as shown in **Figure** 1. Once you have this dimension, calculate the impedance using the outer-to-inner conductor D/d diameter ratio. The characteristic impedance (Z) of a coaxial line is given by:

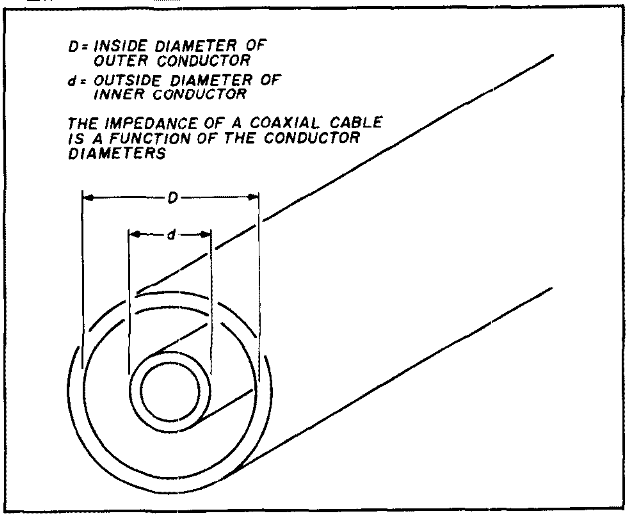
$$Z = (138/\sqrt{\epsilon_r})\log\,(D/d) \qquad (1)$$

where ϵ_r is the relative dielectric constant and d is the inner conductor outside diameter (OD), given in the same unit of measure as the outer conductor The relative dielectric constant for air is extremely close to 1. The impedances for various tubing diameter ratios for copper pipe are shown in **Table 1**. We've listed various combinations of inner and outer conductors to provide a wide impedance selection for any given outer diameter. Impedance selection is especially useful for impedance matching. An impedance-matching section is used frequently in an antenna-phasing harness to match the transmission line impedance to each antenna in an array. The simplest matching section is a quarter-wave transmission line (commonly referred to as a Q matching section), with an impedance equal to the square root of the product of the source and load impedances:

$$Z = Z_S \cdot Z_L \qquad (2)$$

Copper tubing is classified in accordance with the wall thickness, which is based on the required service — as shown in **Table 2**. Household plumbing codes usually require type L pipe. For heating systems, where the same water is recirculated through baseboard heaters, thinner wall type M pipe is acceptable. The outside diameter (OD) of copper tubing is always 1/8" greater than the nominal size. For instance, the chart in **Table** 1 shows that 1/4" tubing has an OD of 3/8" and

FIGURE 1



Anatomy of a transmission line. D = inside diameter of outer conductor; d = outside diameter of inner conductor. The impedance of a coaxial cable is a function of the conductor diameters.

TABLE 1

INNER CONDUCTOR (COPPER)

OUTER CONDUCTOR (COPPER)

		Size	1/4"	3/8"	1/2"	5/8"	3/4"	1"	1-1/4"	1-1/2"	2"	2-1/2"
Size	Type	ID\OD	0.375	0.500	0.625	0.750	0.875	1.125	1.375	1.625	2.125	2.625
3/4"	K	0.745	41.1	23.9	10.5	-	-	-	-	-	-	-
	L	0.785	44.3	27.0	13.7	2.7	-	-	-	-	-	-
1"	K	0.995	58.5	41.2	27.9	16.9	7.7	-	-	-	-	-
	L	1.025	60.3	43.0	29.7	18.7	9.5	-	-	-	-	-
1-1/4"	K	1.245	71.9	54.7	41.3	30.4	21.1	6.1	-	-	-	-
	L	1.265	72.9	55.6	42.3	31.3	22.1	7.0	-	-	-	-
	M	1.291	74.1	56.8	43.5	32.6	23.3	8.3	-	-	-	-
	DWV	1.295	74.3	57.0	43.7	32.7	23.5	8.4	-	-	-	-
1-1/2"	K	1.481	82.3	65.1	51.7	40.8	31.5	16.5	4.5	-	-	-
	L	1.505	83.3	66.0	52.7	41.7	32.5	17.4	5.4	-	-	-
	M	1.527	84.2	66.9	53.5	42.6	33.4	18.3	6.3	-	-	-
	DWV	1.541	84.7	67.5	54.1	43.2	33.9	18.9	6.8	-	-	-
2"	K	1.959	99.1	81.8	68.5	57.5	48.3	33.2	21.2	11.2	-	-
	L	1.985	99.9	82.6	82.6	58.3	49.1	34.0	22.0	12.0	-	-
	M	2.009	100.6	83.3	70.0	59.1	49.8	34.8	22.7	12.7	-	-
	DWV	2.041	101.5	84.3	70.9	60.0	50.8	35.7	23.7	13.7	-	-
2-1/2"	K	2.435	112.1	94.9	81.5	70.6	61.3	46.3	34.3	24.2	8.2	-
	L	2.465	112.8	95.6	82.2	71.3	62.1	47.0	35.0	25.0	8.9	-
	M	2.495	113.6	96.3	83.0	72.0	62.8	47.7	35.7	25.7	9.6	-
3"	K	2.907	122.7	105.5	92.1	81.2	72.0	56.9	44.9	34.9	18.8	6.1
	L	2.945	123.5	106.3	92.9	82.0	72.7	57.7	45.7	35.6	19.6	6.9
	M	2.981	124.2	107.0	93.6	82.7	73.5	58.4	46.4	36.4	20.3	7.6
	DWV	3.035	125.3	108.1	94.7	83.8	74.5	59.5	47.5	37.4	21.4	8.7
4"	K	3.857	139.7	122.4	109.1	98.1	88.9	73.8	61.8	51.8	35.7	23.1
	L	3.905	140.4	123.2	109.8	98.9	89.7	74.6	62.6	52.6	36.5	23.8
	M	3.935	140.9	123.6	110.3	99.3	90.1	75.0	63.0	53.0	36.9	24.3
	DWV	4.009	142.0	124.8	111.4	100.5	91.2	76.2	64.1	54.1	38.0	25.4
5"	K	4.805	152.9	135.6	122.2	111.3	102.1	87.0	75.0	65.0	48.9	36.2
	L	4.875	153.7	136.5	123.1	112.2	102.9	87.9	75.9	65.8	49.8	37.1
	M	4.907	154.1	136.9	123.5	112.6	103.3	88.3	76.3	66.2	50.2	37.5
6"	K	5.741	163.5	146.3	132.9	122.0	112.7	97.7	85.7	75.6	59.6	46.9
	L	5.845	164.6	147.4	134.0	123.1	113.8	98.8	86.7	76.7	60.6	48.0
	M	5.881	165.0	147.7	134.4	123.4	114.2	99.1	87.1	77.1	61.0	48.3
	DWV	5.959	165.8	148.5	135.1	124.2	115.0	100.0	87.9	77.9	61.8	49.1

Characteristic impedance of rigid air line using large copper pipe. (Impedance values given in ohms.)

TABLE 2

Copper tubing classified by wall thickness and required service.

Type	Wall thickness	Service
K	Heaviest wall	Underground
L	Medium wall	General plumbing and heating
M	Thin wall	Sanitary drainage and heating
DWV	Thinnest wall	Drainage, waste, vent

2-1/2" tubing has an OD of 2-5/8". The outside diameter of copper tubing must always be constant for any type within a nominal pipe size (in spite of wall thickness variations), so that solder fittings will be common to each type of tubing.

When you use copper pipe with less than 1" outside diameter, it's desirable to use brass tubing for the inner conductor. Look for it in local hobby shops. The values of impedance are shown in **Table 3**.

As the outside diameter of a coaxial line is increased, the attenuation decreases. There is, however, a limit to the allowable pipe size that can be used for the outer conductor. The minimum attenuation point[7] of a coaxial transmission line occurs at an outer-to-inner conductor diameter D/d ratio of 3.591, which is 76.6 ohms for an air line. Because the attentu-

TABLE 3

INNER CONDUCTOR (BRASS)

			Size	3/32"	5/32"	7/32"	1/4"	9/32"	5/16"	11/32"	3/8"	13/32"	7/16"	15/32"	1/2"	17/32"	9/16"	19/32"	5/8"
	Size	Type	ID\OD	0.094	0.156	0.219	0.250	0.281	0.313	0.344	0.375	0.406	0.438	0.469	0.500	0.531	0.563	0.594	0.625
	5/32"	B	0.125	17.2	-	-	-	-	-	-	-	-	-	-	-	-	-	-	-
	7/32"	B	0.189	42.0	11.5	-	-	-	-	-	-	-	-	-	-	-	-	-	-
	1/4"	B	0.221	51.4	20.9	0.5	-	-	-	-	-	-	-	-	-	-	-	-	-
	9/32"	B	0.253	59.5	29.0	8.6	-0.7	-	-	-	-	-	-	-	-	-	-	-	-
	5/16"	B	0.285	66.6	36.1	15.8	7.8	0.8	-	-	-	-	-	-	-	-	-	-	-
	11/32"	B	0.321	73.8	43.2	22.9	15.0	8.0	1.5	-	-	-	-	-	-	-	-	-	-
	3/8"	B	0.347	78.4	47.9	27.6	19.6	12.6	6.2	0.5	-	-	-	-	-	-	-	-	-
	13/32"	B	0.377	83.4	52.9	32.6	24.6	17.6	11.1	5.5	0.3	-	-	-	-	-	-	-	-
	7/16"	B	0.408	88.1	57.6	37.3	29.4	22.3	15.9	10.2	5.1	0.3	-	-	-	-	-	-	-
	15/32"	B	0.442	92.9	62.4	42.1	34.2	27.1	20.7	15.0	9.9	5.1	0.5	-	-	-	-	-	-
	1/2"	B	0.471	96.7	66.2	45.9	38.0	31.0	24.5	18.8	13.7	8.9	4.4	0.3	-	-	-	-	-
	17/32"	B	0.503	100.7	70.2	49.8	41.9	34.9	28.4	22.8	17.6	12.8	8.3	4.2	-	-	-	-	-
OUTER	9/16"	B	0.536	104.5	74.0	53.6	45.7	38.7	32.2	26.6	21.4	16.6	12.1	8.0	4.2	0.6	-	-	-
CONDUCTOR	19/32"	B	0.565	107.7	77.1	56.8	48.9	41.9	35.4	29.7	24.6	1.8	15.3	11.2	7.3	3.7	0.2	-	-
	5/8"	B	0.598	110.9	80.5	60.2	52.3	45.3	38.8	33.1	28.0	23.2	18.7	14.6	10.7	7.1	3.6	0.4	-
	3/4"	K	0.745	124.1	93.7	73.4	65.4	58.4	52.0	46.3	41.1	36.4	31.8	27.7	23.9	20.3	16.8	13.6	10.5
		L	0.785	127.2	96.8	76.5	68.6	61.6	55.1	49.4	44.3	39.5	35.0	30.9	27.0	23.4	20.0	16.7	13.7
	1"	K	0.995	141.4	111.0	90.7	82.8	75.8	69.3	63.7	58.5	53.7	49.2	45.1	41.2	37.6	34.2	30.9	27.9
		L	1.025	143.2	112.8	92.5	84.6	77.6	71.1	65.4	60.3	55.5	51.0	46.9	43.0	39.4	36.0	32.7	29.6
		K	1.245	154.8	124.5	104.2	96.2	89.2	82.7	77.1	71.9	67.2	62.6	58.5	54.7	51.1	47.6	44.4	41.3
	1-1/4"	L	1.265	155.8	125.4	105.1	97.2	90.2	83.7	78.0	72.9	68.1	63.6	59.5	55.6	52.0	48.6	45.3	42.3
		M	1.291	157.0	126.7	106.3	98.4	91.4	84.9	79.3	74.1	69.3	64.8	60.7	56.8	53.2	49.8	46.5	43.5
		DWV	1.295	157.2	126.8	106.5	98.6	91.6	85.1	79.4	74.3	69.5	65.0	60.9	57.0	53.4	50.0	46.7	43.7
		K	1.481	165.2	134.9	114.6	106.6	99.6	93.2	87.5	82.3	77.6	73.0	68.9	65.1	61.4	58.0	54.8	51.7
	1-1/2"	L	1.505	166.2	135.8	115.5	107.6	100.6	94.1	88.5	83.3	78.5	74.0	69.9	66.0	62.4	59.0	55.7	52.7
		M	1.527	167.1	136.7	116.4	108.5	101.4	95.0	89.3	84.3	79.4	74.8	70.7	66.9	63.3	59.8	56.6	53.5
		DWV	1.541	167.6	137.3	116.9	109.0	102.0	95.5	89.9	84.7	79.9	75.4	71.3	67.5	63.9	60.4	57.2	54.1

B = Brass K, L, M, DWV = Copper

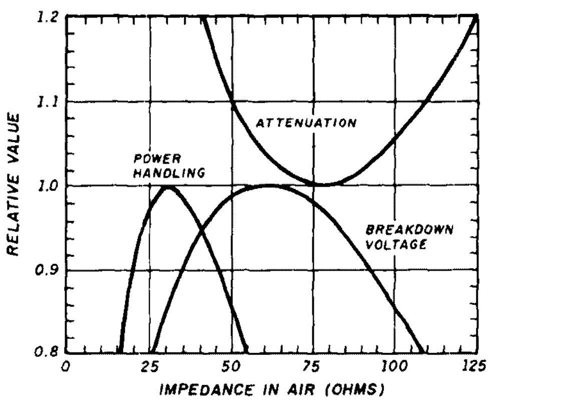
Characteristic impedance of rigid line using brass center conductor tubing. (Impedance values given in ohms.)

ation function is a broad one (see **Figure 2**), the attenutation goes up by a factor of only 1/2 percent at D/d ratios of 3.2 (Z=69.7 ohms) and 4.1 (Z≈84.6 ohms). Thus the conductor diameter ratios need not be exact. The relative attentuation increases by only 5 percent at widely varying *D/d* ratios of 2.6 (Z=57.3 ohms) and 5.2 (Z=98.8 ohms). The best impedances of several criteria for coaxial line are given in **Table 4**, where the minimum attentuation and maximum power are shown for copper pipe using air as a dielectric.

After consulting these impedance tables, you can quickly sort through your junkbox for the pipe diameter closest to optimum. For example, the 2-meter cavity duplexers constructed by John Bilodeau[4] and duplicated by Stewart Gurske[5] used 4″ type DWV copper drain pipe with a 1-3/8″ OD copper tube (a 1-1/4″ size copper pipe has an outside diameter of 1-3/8″) as the center conductor. Dave Baxter[8] constructed a single cavity filter for his 2-meter repeater receiver using the same conductor ratio. The cavity's Q could be improved with the identical 4″ pipe by simply dropping down to a 1″ nominal diameter copper pipe for the center conductor. According to **Table 1**, the characteristic impedance using a 1″ copper pipe center conductor is 76 ohms. This is approximately the minimum attenuation point for a coaxial structure. The impedance of a coaxial line with a 1-1/4″ center conductor is approximately 64 ohms. The relative loss for an impedance of 64 ohms is about 2 percent higher than that of the theoretical optimum value of 76.6 ohms. The unloaded Q of a copper quarter-wave coaxial cavity, designed for an optimum impedance of 76 ohms, is approximated as:

$$Q_u \cong 74D\sqrt{f} \tag{3}$$

FIGURE 2



The attenuation, breakdown voltage, and power handling capability for rigid lines is a function of the D/d ratio.

TABLE 4

Optimized air line copper transmission line parameters.

Cable parameter	Conductor ratio (D/d)	Characteristic impedance
Minimum attenuation	3.591	76.6 ohm
Maximum voltage	2.718	59.9 ohm
Maximum power	1.649	30.0 ohm

TABLE 5

Relative conductivity of commonly available material.

Metal	Relative Conductivity
Silver	1.03
Copper	1.00
Gold	0.84
Aluminum	0.78
Brass	0.48

TABLE 6

Commercial rigid air line copper transmission line dimensions (in inches).

Nominal rigid cable size	Inner conductor OD	Inner conductor ID	Outer conductor OD	Outer conductor ID
7/8 inch (50 ohm)	0.341	0.291	0.875	0.785
1-5/8 inch (50 ohm)	0.664	0.588	1.625	1.527
3-1/8 inch (50 ohm)	1.315	1.231	3.125	3.027
6-1/8 inch (50 ohm)	2.600	2.520	6.125	5.981
6-1/8 inch (75 ohm)	1.711	1.631	6.125	5.981

where D is the inside diameter of the outer conductor in inches and f is the operating frequency in megahertz. If you use other materials, the unloaded Q will vary in direct proportion to the relative resistivity of the metal, as shown in Table 5. Therefore, a brass coaxial cavity will have about one-half the Q of a similar copper cavity. A silver-plated cavity will show a 3 percent increase in unloaded Q.

Large, rigid coaxial transmission lines used for high-power radio and television stations are generally designed with 50-ohm air line. You must pressurize the lines slightly with dry air or nitrogen to prevent moisture condensation. This impedance is a compromise value between power handling capability, voltage breakdown, and attenuation. The attenuation is about 10 percent higher than that of a transmission line with a minimum attenuation impedance of 76.6 ohms. Verify this by checking radio/TV transmitter rigid air line catalogs for 6-1/8" commercial transmission line, where the line is fabricated for both 50 and 75 ohms. Table 6 lists the sizes of the copper pipe used in these installations in case you want to duplicate a high-power, low-loss installation. The 7/8" commercial air line uses common 3/4" type L copper tubing for the other conductor; the 1-5/8" rigid line uses 1-1/2" type M copper tubing. hr

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SIMPLE EEPROM PROGRAMMER

By George Swindell, Jr., WA2IHE, 538 Griscom Drive, Deptford, New Jersey 08096

Anyone who deals with computerized equipment will, sooner or later, come up against Read Only Memory (ROM) ICs. Nowadays, Electrically-Erasable Programmable Read-Only Memory (EEPROM) is replacing the older Ultra Violet-Erasable Programable Read-Only Memory (EPROM) and many of the new ICs are "drop-in compatible" with older types. For example, the 2Kx8 2816 used here matches the popular 2716 EPROM pin for pin. Except during programming, the 2816 needs only a single 5-volt supply.

The EEPROM is a significant improvement over the ultra violet-erasable device. First, it can be erased quickly without an expensive UV source (or waiting for a long, sunny day, hi). Second, single bytes can be erased selectively without disturbing the rest of the ROM's contents.

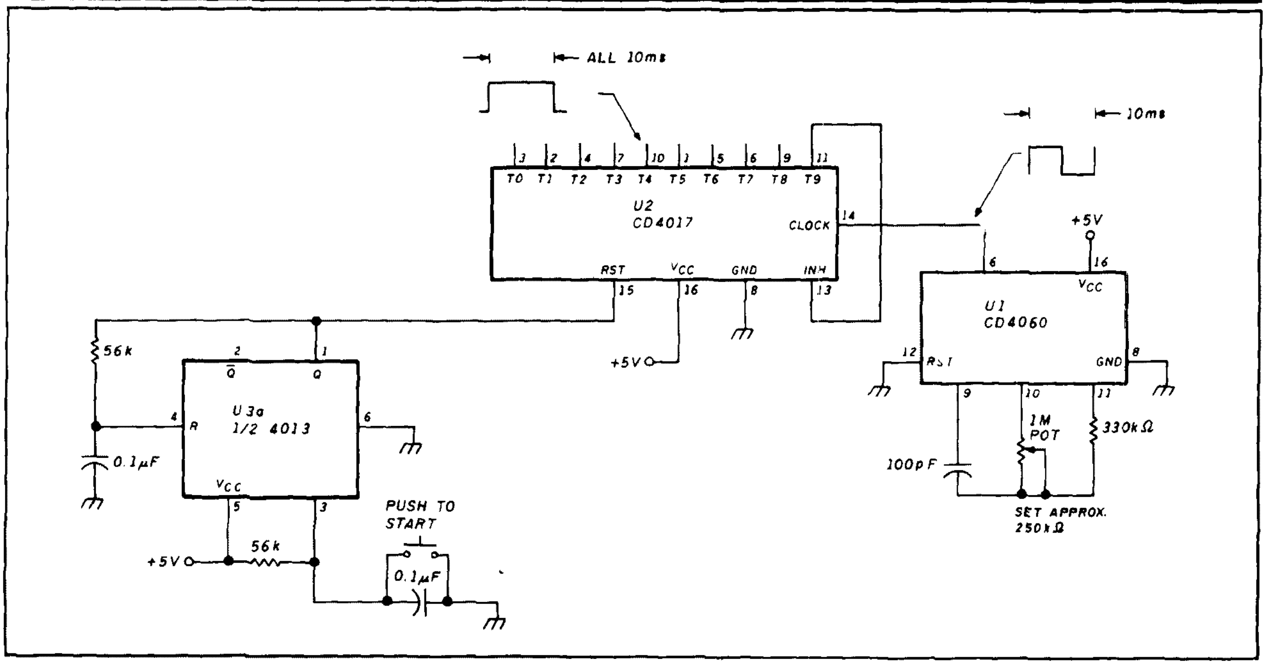
This EEPROM programmer lets you do both erasing and simple, manual programming. It's described as a "stand-alone" device assembled on a prototyping block, but it can just as easily be incorporated into your equipment using the EEPROM. As shown here, the programmer can't be computer driven or operated automatically. You can add these features, but it will cost more in terms of both time and money.

The electrical characteristics for programming show only two *critical* parameters. The programming voltage (V_{pp}) write pulse duration must be between 9 and 15 ms, and the V_{pp} rise time must be between 0.45 and 0.75 ms. Because all other timing parameters are given as minima expressed in nanoseconds or microseconds, I timed everything around the V_{pp} pulse width of 10 ms to keep things simple.

How it works

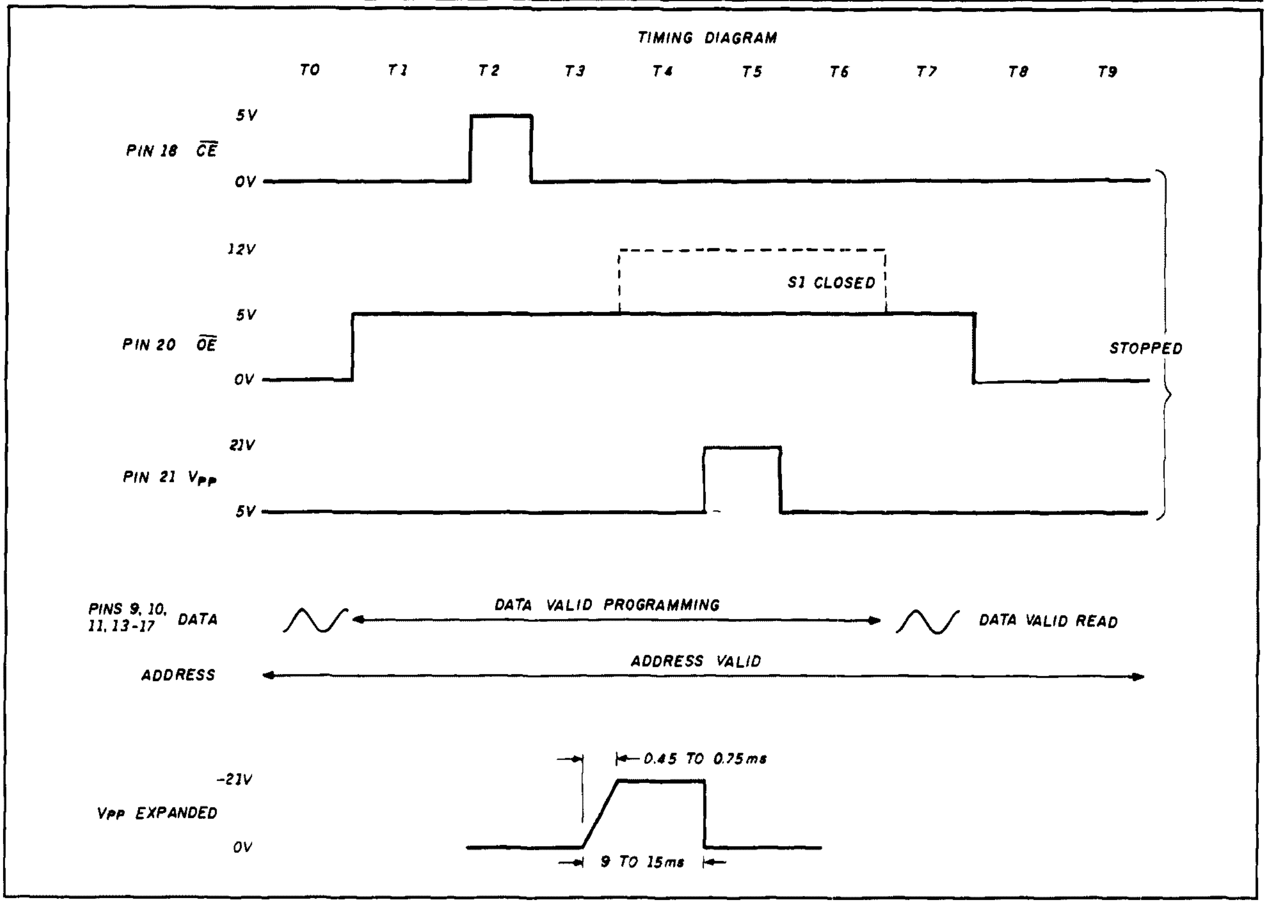
The 4060 oscillator/pulse generator (*U1*) in **Figure 1** is an oscillator/divider combination. The 1-meg pot adjusts the oscillator frequency to 12.8 kHz. This is divided by 128 to

FIGURE 1



Oscillator/pulse generator.

FIGURE 2



Timing diagram.

give a 100-Hz output at pin 6, pulsing the input of pin 14 of the CD4017 (U2), a decade counter which produces sequential 10-ms pulses from the 100-Hz clock input. The pulses appear in order from T0 to T9. When output T9 goes high it activates inhibit, pin 13, stopping all activity. When you press the START switch it resets the 4017, causing T9 to go low and the program to run from T0 to T9 again. Flip-flop U3A is a debounce circuit.

Flip-flop U3B has already been reset when the counter arrives at T9, causing the outputs of the 4503 analog multiplexer/demultiplexers (U4 and U5) to float. The inputs to the 4049 hex inverting buffers (U6 and U7) from the EEPROM socket cause the LEDs to display the value of data being read from the EEPROM. Refer to the timing diagram in **Figure 2**. When you push START, T0 goes high for 10 ms and then goes low. Next, T1 goes high and sets flip flops U3B and U8B causing the value of the data thumbwheel switches (shown in **Figure 3**) to be applied to the I/O pins of the EEPROM. Also, $\overline{OE}$ (**Figure 4**) pin 20 goes to 5 volts. T2 then applies a 10-ms pulse to pin 18 of the EEPROM. T4 sets flip-flop U8A to apply a 12-volt pulse to pin 20. This pulse is used for chip erase. Leave the 26-volt switch S1 open for a byte erase and the pulse will remain at 5 volts, as indicated by the solid section of the timing diagram in **Figure 2**, EEPROM $\overline{OE}$. The LM317 voltage regulator (U10) normally supplies 5 volts to the V_{pp} pin. The 1 and 4.4-k resistors are in parallel because T5 is still low. The bottom of the 1-k resistor is effectively at ground. When T5 goes high, LM317 voltage regulator U10 is allowed to rise to 21 volts. The time constant of the 1-k resistor and 0.02-μF capacitor produces a rise time of about 0.6 ms, which should improve device reliability. After T5 the process reverses itself, and the circuit comes to a stop at T9. The address pins are connected directly to the thumbwheel switches in **Figure 5** and you must enter the correct address before any byte erase/write takes place. The resistors at the bottom of the data and address switches terminate the CMOS inputs of IC 4503 and the address pins A0 to A10.

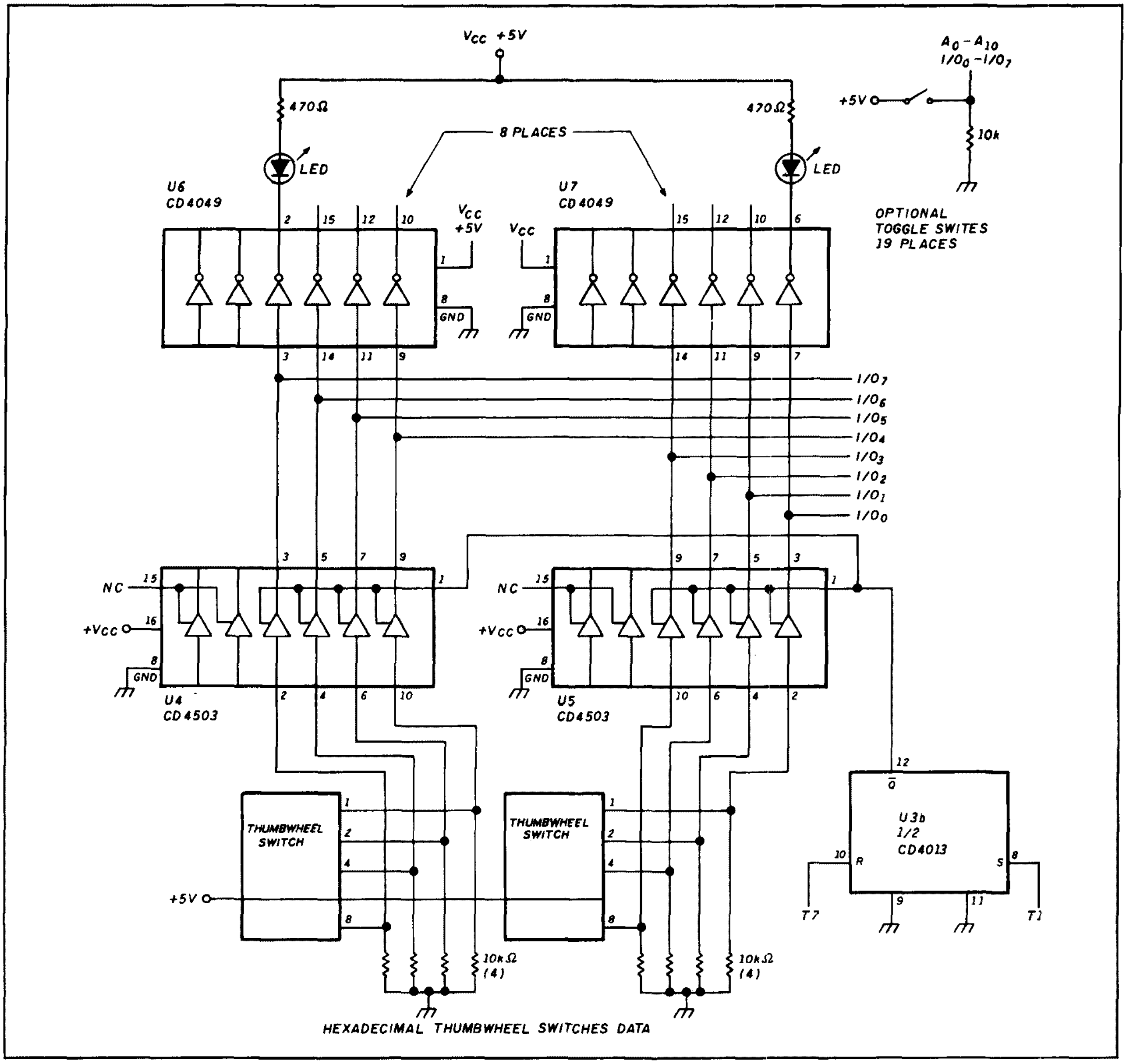
Power supply

I didn't build a power supply for this project. I have several 1 to 15-volt regulated supplies with floating outputs. I connected two of them in series to obtain 26 volts. The current drain is very small, around 20 mA maximum. The 5-volt supply uses about 200 mA.

Adjustments

Temporarily disconnect pin 11 of the 4017 and ground pin 13. This causes the counter to run constantly. Apply

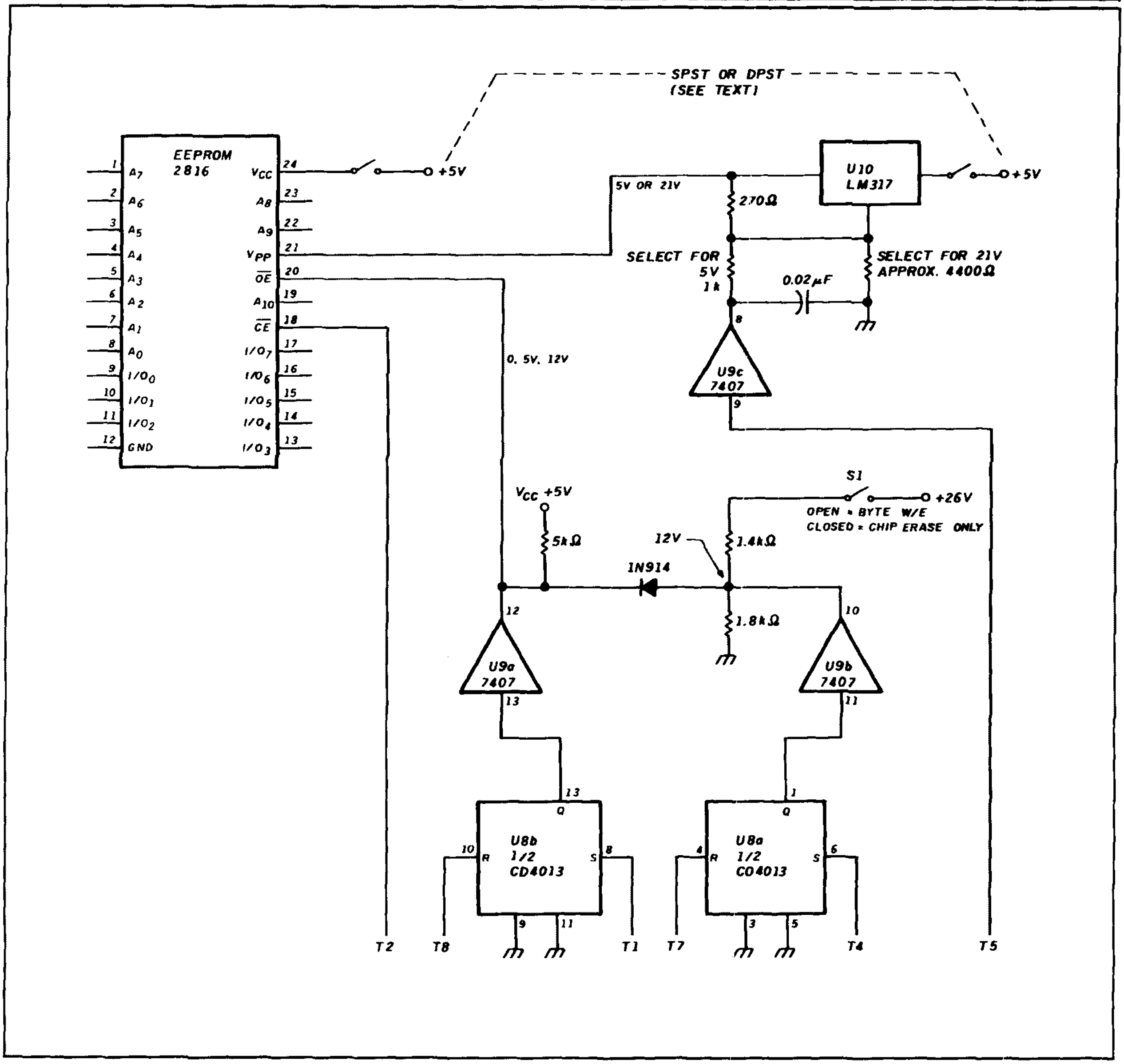
FIGURE 3



Data setup and readout.

power to all circuits with the EEPROM removed from its socket. Using a scope, sync on T0 (U2 pin 3) and adjust the 1-meg pot on IC 4060 for a pulse width of 10 ms. Keep sync on T0 and adjust the scope sweep speed for 10 ms/cm. T0 to T9 are now each represented by centimeters across the screen. Refer to the timing diagram and observe $\overline{CE}$, pin 18 of the EEPROM socket. The pulse should be at T2. $\overline{OE}$ pin 20 should appear as a 5-volt pulse. Closing S1 causes a 9 to 15-volt pulse to be superimposed on the 5-volt pulse as indicated by the dotted lines. The 1.4 and 1.8-k resistors establish this voltage. V_{pp} pin 21 should start out at 5 volts (adjust the 1-k resistor if necessary). The pulse then goes to 21 volts. Adjust the 4.4-k resistor for 21 volts $\pm$1 volt. Next, expand the trace speed to x10 and observe the rise time of the 21-volt pulse; it should be between 0.45 and 0.75 ms. Adjust the 0.02-μF capacitor if necessary. The pulse should be between 9 and 15 ms long, so go back and tweek the 1-meg pot if necessary. Now return to normal sweep. Check the I/O pins by observing the waveforms. If you touch the probe tip you should see a 60-cycle hum, indicating that the outputs of IC 4503 are floating. The solid line from T1 to T7 should be either low or high depending on the setting of the data switches. Next, check each address pin to see if you're getting good highs and lows. Remove the ground on IC 4017 pin 13 and reconnect pin 11; the counter should stop. Push START; one cycle should pass

FIGURE 4



Programming control.

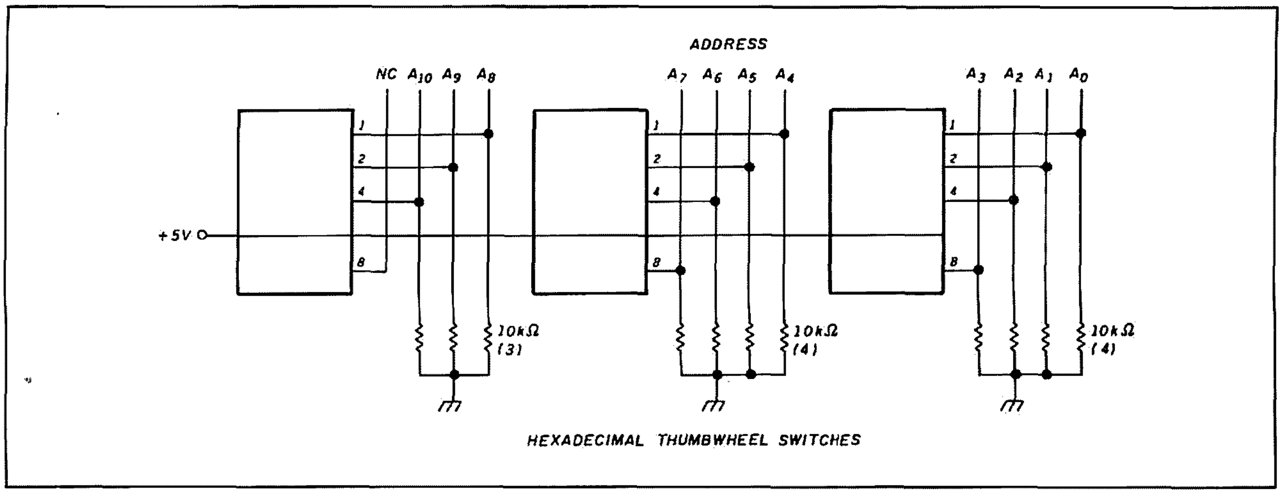
and stop at T9. If two or more cycles pass, you may have to adjust the debounce flip-flop for a longer time constant.

Using the circuit

This circuit can power up at any point from T0 to T9, applying unwanted pulses. To prevent this, I've placed a toggle switch to inhibit the EEPROM's V_{CC} 5-volt and 26-volt, which inhibits V_{PP}. First insert the 2816 into its socket and turn on the main power to the rest of the ICs. Push START once or twice to reset all flip-flops. If you're using an SPST switch, apply 5-volt V_{CC} power first — then apply the 26-volt power. If you have a DPST switch, both V_{CC} and V_{PP} will be applied simultaneously. Power down in the reverse order when removing the 2816 from its socket.

The EEPROM should be in read mode after power up. A new EEPROM should show all highs (LEDs lit) for the erased condition. The LEDs and LED drivers can be eliminated if you don't mind touching each I/O pin with a VOM to determine the logic level. Place the address switches at 000 or any desired address. If the data byte you're looking at has been erased already (all highs), enter the value of data you want at that address and push START. The LEDs should show the same value entered on your data switches because you're reading the 2816 at T9. If data is already present and you want to change it, first erase that location by writing all "1"s (FF_H), then enter the correct data and push START again. To erase the entire 2K of memory, close S1 and push START. The entire chip will be erased. I've

FIGURE 5



Address selection.

found that the address and data settings aren't important when you're doing a chip erase using the SEEQ device. However, another 2816 by EXEL requires the data value FF for a chip erase as well as a byte erase. I suggest you obtain data sheets for any EEPROMs you buy to determine if there are any special procedures to follow.

Conclusion

I built the programmer on a protoboard so I could change the circuitry to accommodate other PROMs. You may want to try using the EEPROM to decode binary values into seven-segment readouts including the characters A to F. Circuits with the 2716 UV EPROM can be modified using the 2816. You can program eight outputs with varying pulse widths to control other circuits by hooking a binary counter up to the address pins. I used three EEPROMs with their addresses in parallel for a total of 24 control lines. Those 24 lines let me program a single chip microcomputer with its own UV EPROM. I have since replaced the EEPROMs with a single chip microcomputer which programs other PROMs. You can purchase all of the ICs used in this project from Jameco, 1355 Shoreway Road, Belmont, California 94007. hr

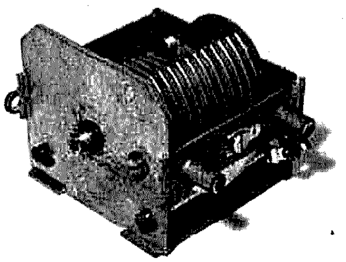
NEAR LINEAR TUNING
WITH DUAL ECCENTRIC PULLEYS

Straight line variable caps cover a large tuning range

By John Pivnichny, N2DCH, 3824 Pembrooke Lane, Vestal, New York 13850

Extraordinarily linear tuning is possible with ordinary variable capacitors if you use some new techniques. As my earlier articles[1,2] point out, all modern Amateur transceivers have very linear frequency dials. Unless they use a digital PLL technique, commercial products use precision, specially made, gear-driven variable capacitors not available to the homebrewer. This article describes a dual eccentric pulley approach, which allows common straight line capacitance variables to be used over fairly large frequency ratios.

PHOTO A



Ordinary capacitor.

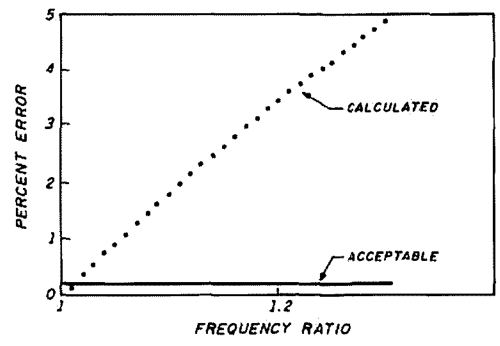
Variable capacitors

Ordinary variable capacitors have semicircular rotor plates and uniform change in capacitance as the shaft is rotated through 180 degrees from full mesh to completely open (see Photo A). This is known as a straight line capacitance variable. When used as the main tuning element, it produces a frequency change that compresses the scale at the high frequency end.

Shape of rotor plates

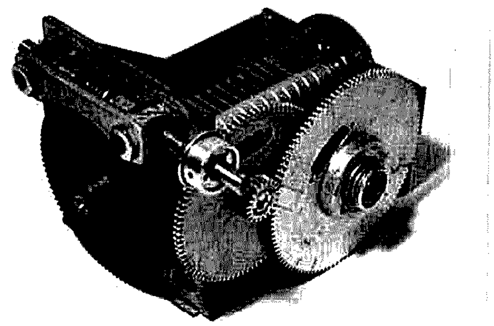
If the capacitor rotor plates are specially shaped, this dial compression can be spread out. In general, the radius of the rotor plates must be reduced at the minimum capacitance end so that each degree of rotation causes less capacitance change. With correctly shaped rotor plates, an equally spaced linear frequency scale results. The plates are referred to as "straight line frequency variables." Capacitors constructed this way are usually not available or very expensive. One exception is the WWII surplus capacitor described in Reference 2 and shown in Photo B.

FIGURE 1



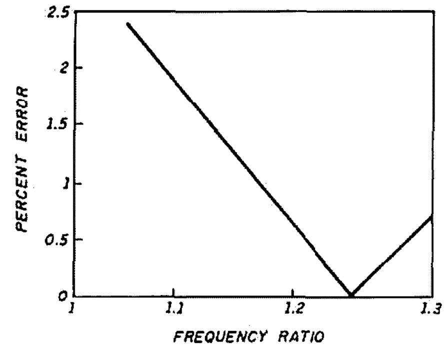
Percent error for straight line capacitance variable.

PHOTO B



World War II capacitor.

FIGURE 2



Percent error for special World War II capacitor.

Worst-case dial error

Equations in the appendix of Reference 1 show how to calculate the degree of error a linear dial will have when a straight line capacitance variable is used. The frequency at the dial center will be low if the end points are set exactly. Note that the worst-case error can be reduced to one-half by setting the frequency at both ends of the dial high by one-half the center dial error. This error is shown in the graph in Figure 1, plotted as a percentage of full-range error for various frequency tuning ratios. The range is defined as high frequency minus low frequency, and the errors have been reduced to one-half.

The curve is very steep; the error is unsatisfactory for all but very low tuning ratios and narrow ranges. I consider an error of ± 1 kHz in a 500-kHz range satisfactory and an error much beyond that as excessive. For example, the area below the dotted line in Figure 1 represents the acceptable region.

Low satisfactory tuning ratios force you to select fairly high frequencies to cover a reasonable range. In Reference 1 I chose 11.5 to 11.75 MHz to cover a 250-kHz range with a ± 2-kHz worst-case error. That let me receive 3.5 to 3.75 MHz with an IF of 8 MHz.

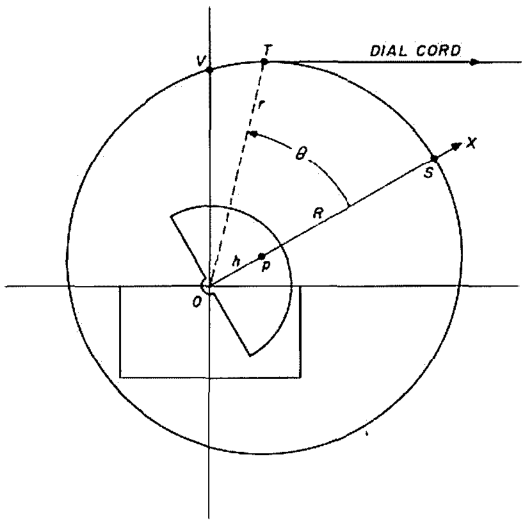
Comparison with the curve of Figure 2 for the specially constructed WWII capacitor[2] shows that very low errors are possible for a 1.25:1 frequency ratio. Such capacitors are usable for a single ratio. However, the technique described here can be used for many frequency ratios.

Eccentric pulleys

One way to reduce the capacitance change at the high frequency end is to mount an off-center (eccentric) pulley on the shaft and drive it with dial cord. The mounting must be made so the cord winds or unwinds with a maximum radius when the capacitor is at the high frequency or low capacitance end. Check the conceptual sketch in Figure 3 for details.

If this pulley were just the right shape, you'd have the same effect you'd have if the capacitor plates were cut to exactly the right shape for linear frequency response. Such a pulley would be very difficult to make. On the other hand, circular pulleys are easy to make from Plexiglas™ as described in Reference 1. Note, however, that the shaft mounting hole will be off center — you can't use it in the initial turning, filing, and sanding of the rough cut pulley when it's mounted in an electric drill.

FIGURE 3



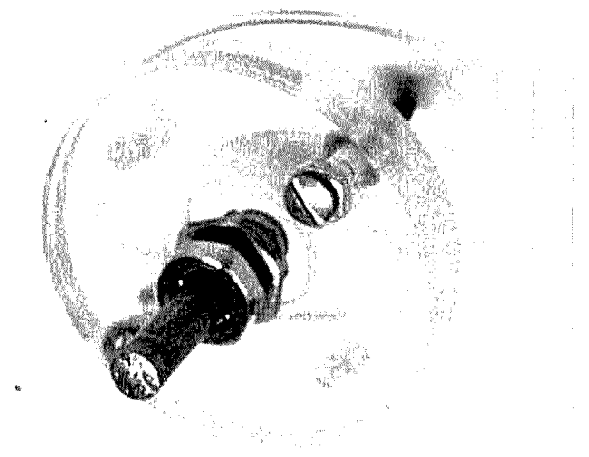
Eccentric pulley.

Instead, the rough cut pulley is mounted on center to a mandrel with machine screws as shown in Photo C. The off-center shaft hole is drilled after the pulley has been completed and removed from the electric drill. I realize that you may wonder how far off an eccentrically mounted circular pulley arrangement will be, and if it will be close enough to the right shape for a linear frequency scale.

Theory

Figure 3 shows a pulley with center P and radius PS of length R. It's mounted on a variable capacitor at point O, which is length h off center. The capacitor and pulley rotate around point O. A dial cord pulls horizontally on the pulley. Note that distance r from tangent point T to center O changes as the pulley rotates through angle θ from 0 degrees (plates fully unmeshed) to 180 degrees (plates fully meshed).

PHOTO C



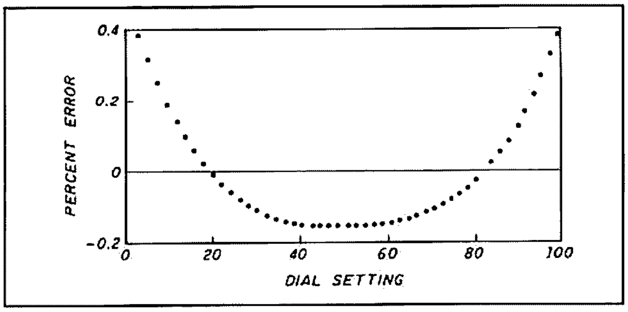
Mandrel.

TABLE 1

Typical calculations for h = 0.5, R = 1, and external capacitance $C_{ext} = C_{max} - C_{min}$.

θ (degrees)	Length TS (inches)	Frequency F (kHz)
0	0	0.1
1.8	0.04711	0.09950
3.6	0.09419	0.09901
5.4	0.14121	0.09853
7.2	0.18815	0.09806
9.0	0.23498	0.09759
10.8	0.28169	0.09713
12.6	0.32825	0.09667
↓	↓	↓
180	3.1416	0.07071

FIGURE 4



Best fit deviation of calculated frequency from true straight line.

Because the distance r changes, the amount of capacitor rotation produced by a given increment of dial cord movement also changes. This is the desired effect. Note that the point of tangency T will coincide with the vertical y axis at $\theta = 0$ degrees and $\theta = 180$ degrees. For other angles in between it moves slightly to the right as shown. The difference in distances OT and OV will be small and are ignored in this analysis. Mathematically, length r is given for any angle θ by the following equation:

$$r = h\left[\cos\theta + \sqrt{\left(\frac{R}{h}\right)^2 - \sin^2\theta}\right] \qquad (1)$$

This formula is derived in the article's appendix. The appendix begins with a formula from an old high-school math textbook.[3] It's not necessary to understand the details of these equations to use this technique in Amateur construction. I've reduced the math to graphical easy-to-use results.

Finding perimeter distance TS

As the dial cord is pulled to the right, it unwinds from the pulley. The unwound length is perimeter distance TS. Assume the dial cord is pulled an equal amount for each increment of change in the main tuning knob and dial. This is what happens with the usual dial and tuning knob arrangement, where the dial is driven by an on-center pulley of approximate radius R/2. That is, the dial rotates through 360 degrees for a 180-degree rotation of the capacitor pulley. If you know distance TS for each angle θ, you'll know the capacitance change for each increment the dial cord is pulled, or (in effect) each increment of dial rotation.

The method for finding distance TS using numerical integration follows. I calculated the distance r from Equation 1 by taking 100 increments of θ from 0 degrees to 180 degrees for a preselected offset h. The perimeter distance traveled through a 1.8-degree increment of θ is equal to:

$$s = 2\pi r \cdot \frac{1}{360} \qquad (2)$$

Each small increment s will be slightly shorter than the previous one. Add these increments of length from 0 to θ degrees for each angle θ. I now have a table of values of length TS for all 100 values of θ. I repeat the process for each offset value h. One such table is shown in the first two columns of Table 1 (for h = 0.5 R). I recommend that you use a personal computer for these calculations.

With the arrangement shown in Figure 3, the capacitance at $\theta = 0$ is equal to the variable capacitor's minimum value C_{min} plus any fixed external capacitance. Call the total:

$$C_o = C_{min} + C_{ext} \qquad (3)$$

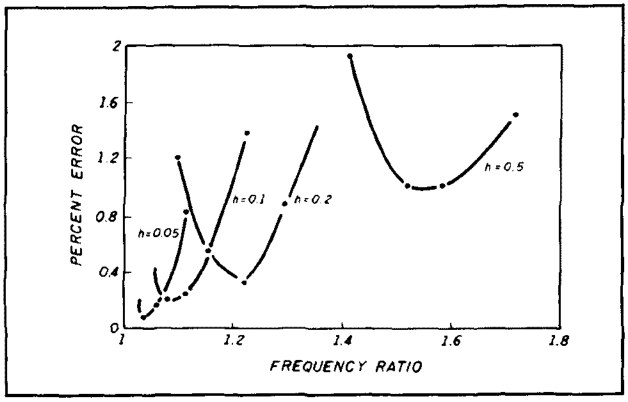
As the capacitor rotates to angle θ degrees, its capacitance increases to the following:

$$CAP = (C_{max} - C_{min}) \times \frac{\theta}{180} \qquad (4)$$

The frequency, however, is given as follows:

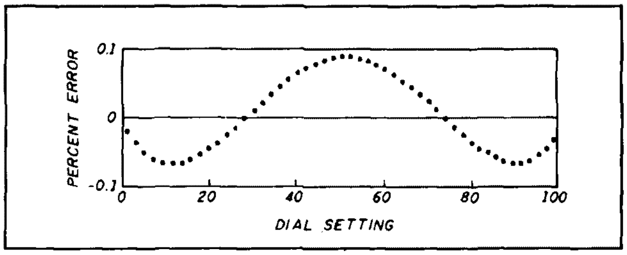
$$F = \frac{1}{2\pi\sqrt{L(C_o + CAP)}} \qquad (5)$$

FIGURE 5



Percent error as a function of frequency ratio for four offset values h. This is the main result of the calculations.

FIGURE 6



Deviation with external capacitance set for minimum error.

Select a value for C_0 (a value of external capacitance). Calculate CAP from **Equation 4** and f from **Equation 5** for each of the 100 values of θ in the table. Include these numbers as a third column in the table.

Frequency linearity

For a particular value of pulley offset h and a particular value of external capacitance C_{ext}, you now have 100 values of dial position TS and 100 corresponding frequencies. You may wonder about the linearity of the resulting frequency if you're using a linear dial like one made from a circular protractor (suggested in **Reference 1**). This question is best answered by drawing a best-fit straight line through the points on a graph of frequency F versus dial position TS. The deviations of the points calculated from this best-fit straight line give the error. A graph of these deviations is shown in **Figure 4**. I prefer to divide the error by the frequency range covered and call this the percent error. Then I plot percent error versus frequency ratio f_{max}/f_{min}.

For each offset value h, you can try a range of fixed external capacitance to see which gives the lowest percent error. Note that frequency range covered, and therefore the frequency ratio, will be different for each value of external capacitance.

Frequency ratios

These data points are plotted in **Figure 5** for values of offset: h = 0.05, 0.1, 0.2, and 0.5. Note that there's a range of values for each h value where the percentage error is quite low. To the left and right the error gradually increases. For larger offsets h, the minimum occurs at larger frequency ratios.

This is to be expected, because a larger frequency ratio requires a larger difference in the capacitance changes at the dial ends than a low ratio does. Larger offsets h produce larger differences in the distance r at the C_{max} and C_{min} ends.

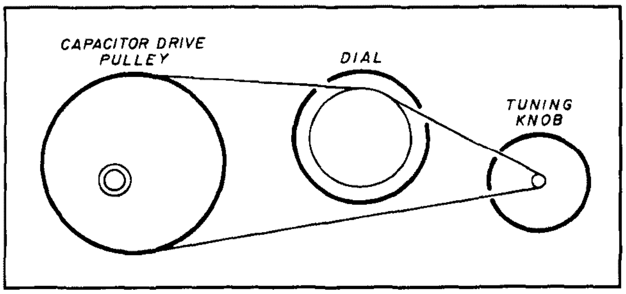
Also note that the minimum percentage error is larger with larger values of h. This is undoubtedly because an off-center circle isn't exactly the right shape for a linear frequency scale. The variation from the correct shape becomes greater at larger frequency ratios.

Nevertheless, quite acceptable percentage errors are produced with the easy-to-build, circular, off-center mounted pulleys. As I mentioned earlier, I consider an error of 0.2 percent, or 1 kHz, in a 500-kHz frequency range dial very acceptable. This is about the limit at which I can set such a dial by eye, if I'm very careful.

Figure 6 shows another plot of frequency differences from a best-fit straight line with an external capacitance set for an error near the minimum. I've included this to illustrate the type of error you would normally experience because you'd usually select a deviation h and external capacitance C_0 for a minimum error.

Figure 5 is the key operating result of all this theory. You can use it to optimize any off-center tuning design. You can also see what effect small differences in mechanical dimensions or external capacitances will have. In general, because the curves are so flat near the minima, minor construction differences won't have much consequence. Small percentage errors are readily achievable as the next section will show.

FIGURE 7



Impossible arrangement.

Practical verification

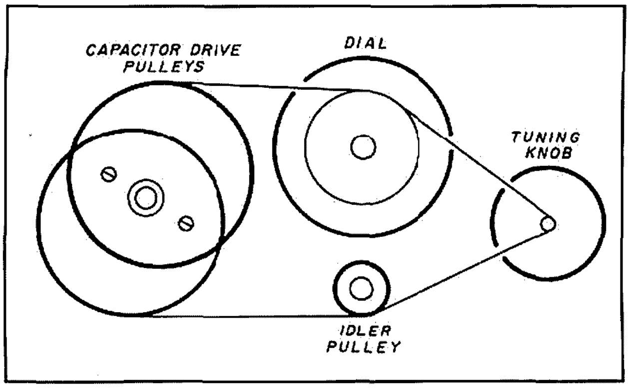
I built an operating model to verify these results. A note on the overall dial stringing is necessary here. Any off-center pulley like the one shown in **Figure 3** requires that the dial cord be returned to the pulley at the same (varying) rate it is removed. Otherwise the cord will become too loose or too tight. The simple sketch in **Figure 7** won't work with this off-center pulley.

Resolve the problem by using a second off-center pulley, offset in the opposite direction. **Figure 8** shows an offset which has been greatly exaggerated. As the main tuning

(continued on page 60)

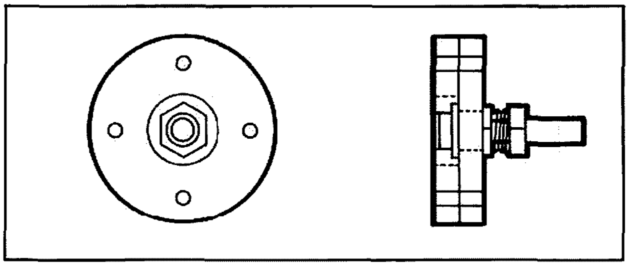
(continued from page 56)

FIGURE 8



Dual eccentric pulley arrangement.

FIGURE 9



Mandrel for pulley construction.

knob is rotated clockwise, the top dial cord comes off the off-center pulley and an equal amount is picked up by the other pulley from the bottom cord. You need a small idler pulley to keep the bottom cord horizontal (approximately). This is the arrangement I selected and built. It's shown in Photos D, E, and F and operates very smoothly.

Pulley details

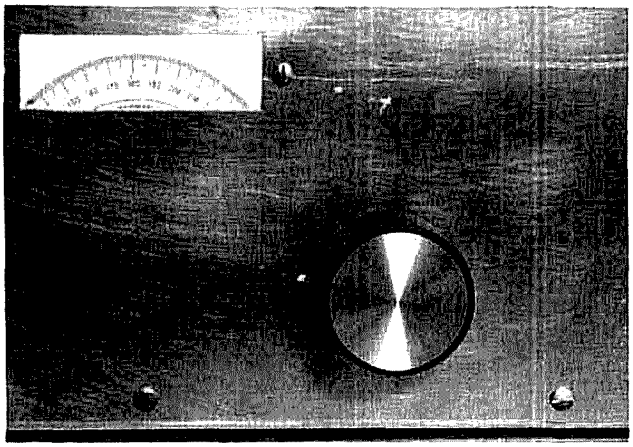
I completed the pulley construction using the method in Reference 1. I also built the mandrel shown in Figure 9 and Photo C using the same techniques. This mandrel holds the large pulleys on center when each is mounted temporarily to the mandrel with four machine screws. After a large pulley is shaped and the groove is cut, it's removed from the mandrel and its mounting hole is drilled off center. My large pulleys have a diameter of 3-1/2 inches. The mounting hole was drilled 0.175 inches off center for $h = 0.1$.

You can do this easily by marking the off-center location with an awl, drilling it with a small 1/16-inch or smaller drill, and increasing the hole diameter with gradually increasing bit sizes. The final one should be 3/8 inch. The second large pulley has its mounting hole cut for 3/4 inch with a hole-cutting attachment, like the one shown in Reference 1. This larger size lets you mount the locking shaft securely. The two pulleys are held together with machine screws. Check the photographs for details.

Pulley sizes

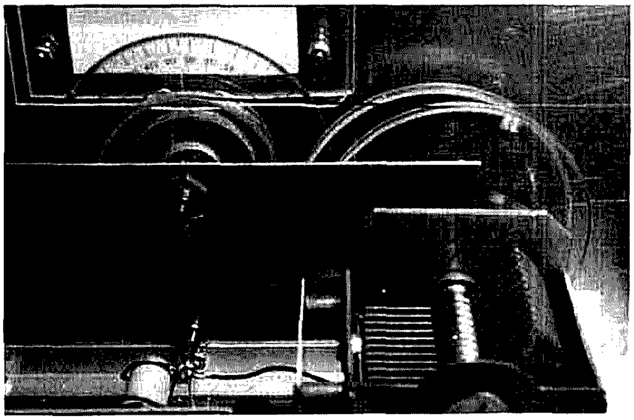
I mounted an aluminum front panel and subpanel on an aluminum bottom plate following the sketches in Reference 1. In this case the dimensions were as follows:
Dial pulley diameter: 2-1/4 inches.
Dial pulley centered 2-5/8 inches above base plate.

PHOTO D



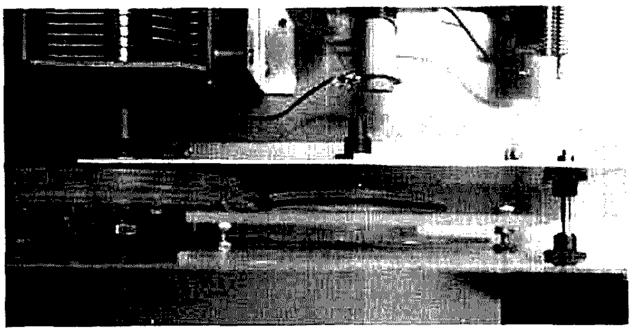
Dial assembly, front view.

PHOTO E



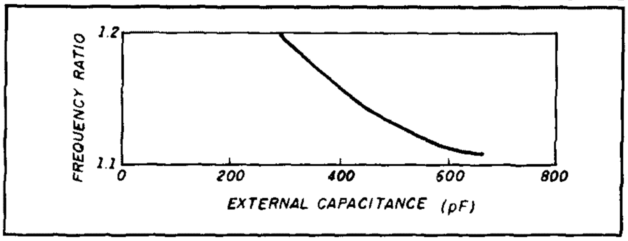
Dial assembly, back view.

PHOTO F



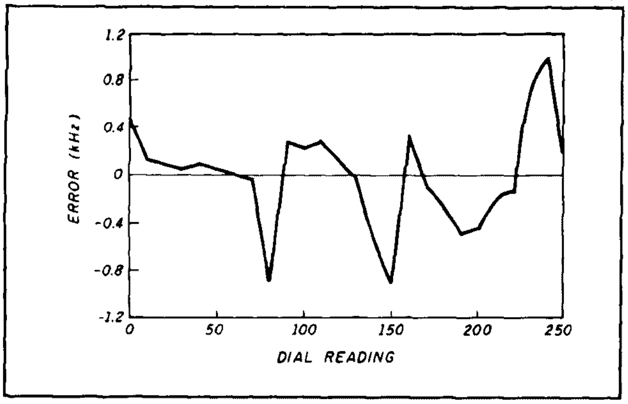
Dial assembly, top view.

FIGURE 10



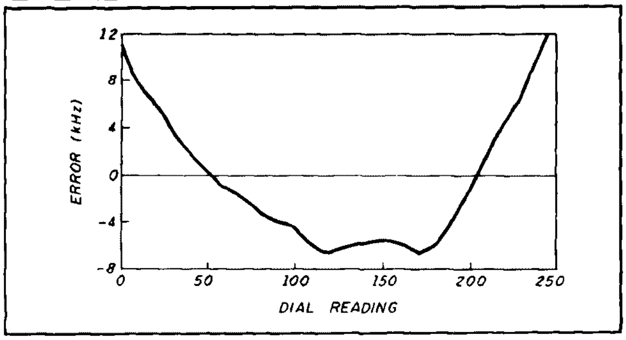
Frequency ratios for various external capacitors.

FIGURE 11



Deviation of measured frequencies with 600-pF external capacitance (2.0 to 2.2 MHz).

FIGURE 12



Deviation of measured frequencies (2.8 to 3.4 MHz; 220 pF external).

Capacitor pulley centered 2 inches above base plate.
Idler pulley diameter: 3/4 inch.
Idler pulley centered 11/16 inch above base plate.

I cut the shafts to length, and once all were operating smoothly I strung the dial cord as shown in **Figure 8**. You should use a small spring at one end to keep the dial cord under tension. The arrangement is shown in **Photos D** (front view), **E** (back view), and **F** (top view).

TABLE 2

Measured results for 5.5 to 6.0-MHz range.

dial	Frequency (kHz)	dial	Frequency (kHz)
0	5500	260	5760
20	5519	280	5780
40	5539	300	5801
60	5559	320	5820
80	5579	340	5839
100	5600	360	5859
120	5620	380	5880
140	5640	400	5899
160	5660	420	5920
180	5680	440	5940
200	5700	460	5961
220	5720	480	5981
240	5741	500	6000

The circuit

It's important to use a high quality capacitor. For my setup, I used the ARC5 oscillator capacitor[1] shown in **Photo A**. **Photo E** has the details. I also used the CA3028 circuit with emitter follower buffer of **Reference 1**. I tried several values of fixed capacitor to establish the value needed for a 1.1:1 frequency ratio. The graph in **Figure 10** shows the measured results. About 600 pF is right for the dial with its h = 0.1.

Frequency readings

With the pulley directions just given, a 250-degree rotation of the circular protractor dial gives slightly less than 180 degrees of capacitor rotation. I took readings every 10 degrees from 0 to 250, then plotted and curve fit them to a straight line. **Figure 11** shows the results. The less than 0.8-kHz error for a 2 to 2.2-MHz range agrees with the calculated results shown in **Figure 5**.

Another check with 220-pF external capacitors is shown in **Figure 12**. The resulting 1-percent error also agrees with **Figure 5**.

Finally, I decided to see how well this off-center apparatus can cover the 5.5 to 6-MHz range. This is the 500-kHz range used in my Kenwood transceiver. I replaced the protractor labeling with a paper disc of new labels from 0 to 500, covering the 0 to 250 degrees on the original labels. I left the outer degree lines visible and indicated 2-kHz markings. Then I mounted an external 140-pF APC trimmer capacitor next to the oscillator coil and changed the fixed capacitors as shown in **Figure 13**.

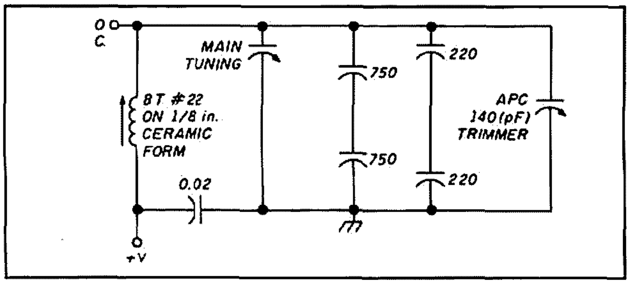
After adjusting the inductor for 5.500 MHz at the low end and the trimmer for 6.000 MHz at the high end, I took frequency counter readings for every 20-kHz position of the dial. **Table 2** shows my results. A 1-kHz maximum error across the dial is acceptable, in my opinion.

Conclusion

I find it practical to perform near linear tuning using dual eccentric pulleys, and know of no other way to get this kind of linearity out of a straight line capacitance variable. The

technique appears to be new, probably because the math required for analysis is quite involved. But by using the main result of Figure 5, you can build practical tuners easily in your home workshop.

FIGURE 13



External capacitors for 5.5 to 6.0-MHz range.

Appendix

To derive Equation 1, start with the general equation of a circle:

$$x^2 + y^2 + Ax + By + C = 0 \qquad \textbf{(6)}$$

This is given in polar coordinate form by Currier[3] on page 291:

$$r^2 + r(A \cos \theta + B \sin \theta) + C = 0 \qquad \textbf{(7)}$$

For the circle in Figure 3, the x axis is as shown and the y axis would be perpendicular to it. Angle θ and distance r are shown. A circle centered at x = h, y = o is given by the following equation on page 240 of Reference 3:

$$(x - h)^2 + y^2 = R^2 \qquad \textbf{(8)}$$

Value R is the distance from P to S, the true radius of the circle. Expanding Equation 8 you get:

$$x^2 + y^2 - 2hx + h^2 - R^2 = 0 \qquad \textbf{(9)}$$

Compared to Equation 6, A = −2h, B = 0, and C = $h^2 - R^2$.

So Equation 7 can be written for the circle of Figure 3:

$$r\pm + r(-2h \cos \theta) + h^2 - R^2 = 0 \qquad \textbf{(10)}$$

Solve for r using the quadratic equation:

$$r = 1/2\ [2h \cos \theta \pm \sqrt{4h^2 \cos^2 \theta - 4(h^2 - R^2)}\,] \qquad \textbf{(11)}$$

Using the well-known identity $1 - \cos^2 \theta = \sin^2 \theta$:

$$r = h\left[\cos \theta \pm \sqrt{\left(\frac{R}{h}\right)^2 - \sin^2 \theta}\,\right] \qquad \textbf{(12)}$$

Rejecting the negative value gives you Equation 1. hr

REFERENCES
1. John R. Pivnichny, N2DCH, "A Homebrew Tuning Dial," *Ham Radio*, December 1988, page 75
2. John R. Pivnichny, N2DCH, "Linear Tuning with a War Surplus Capacitor," *Ham Radio*, June 1989, page 40.
3. C.H. Currier, E.E. Watson, and J.S. Frame, *A Course in General Mathematics*, Macmillan, New York, 1957

INTRODUCTION TO **WAVEFORM GENERATORS** PART 3

Triangle and sawtooth waveform generators

By Joseph J. Carr, K4IPV, P.O. Box 1099, Falls Church, Virginia 22041-0099

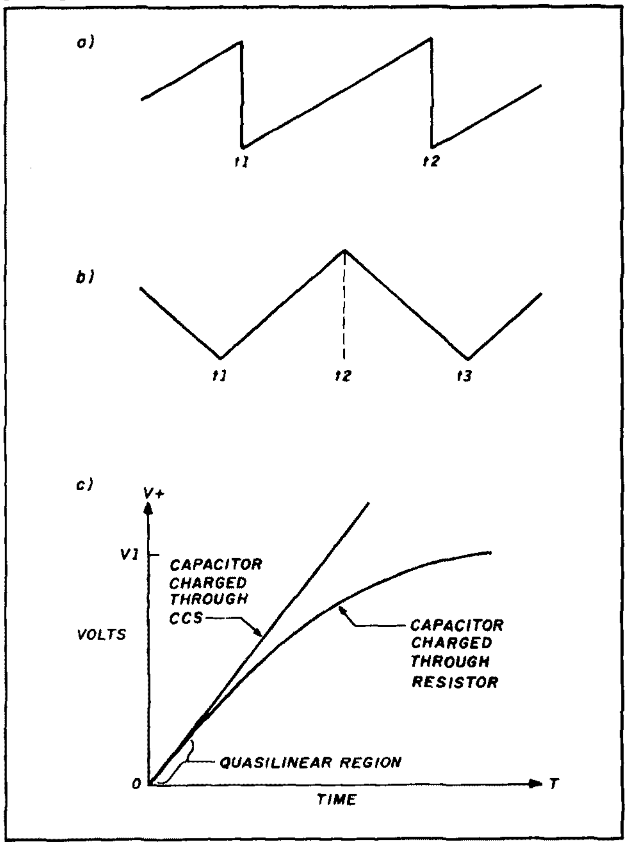
T*riangle* and *sawtooth* waveforms (**Figure** 1) are examples of *periodic ramp* functions. The sawtooth in **Figure 1A** is a single ramp waveform. The voltage begins to rise linearly at time t1. At time t2 the waveform drops abruptly back to zero, where it starts to ramp up linearly again. The sawtooth is usually periodic, although single sweep variants are sometimes used. The period is defined as T (see **Figure 1A**), so the frequency is 1/T.

The triangle waveform in **Figure 1B** is a double ramp. The waveform begins to ramp up linearly at time t1. It reverses direction at time t2 and then ramps downward linearly until time t3. At time t3 the waveform reverses direction again, and begins ramping upwards. The period of the triangle waveform (T) is T1-T3.

Ramp generators are derived from capacitor charging circuits. I discussed the familiar RC charging curve earlier in this series. It's reproduced in simplified form in **Figure 1C**. The RC charging waveform has an exponential shape, so it's not well suited to generating a linear ramp function. There are two approaches to forcing the capacitor charging waveform to be more linear. The first is to limit the charging time to the short quasi-linear segment shown in **Figure 1C**. The ramp obtained isn't very linear, is limited in amplitude to a small fraction of V1, and has a relatively steep slope that may or may not be useful for any given application. A superior method is to charge the capacitor through a *constant current source* (CCS). Using the CCS to charge the capacitor results in the linear ramp shown in **Figure 1C**.

Triangle and sawtooth waveform oscillators create the constant current form of ramp generator by means of a Miller integrator used to charge the capacitor (see **Figure 2A**). When a Miller integrator is driven by a stable reference volt-

FIGURE 1



(A) Example of a sawtooth waveform. Time t1 to t2 represents one cycle. (B) Example of a triangle waveform. Time t1 to t3 represents one cycle. (C) Graph illustrating why a CCS capacitor produces a more linear and uniform ramp.

age, the output is a linearly rising ramp. The ramp voltage (V_o) is:

$$V_o = \frac{V_{ref}}{T} \tag{1}$$

or, because T = RC:

$$V_o = \frac{V_{ref}}{RC} \tag{2}$$

If V_{ref} = +10 volts DC, and the RC time constant is T = RC = 0.001 seconds, the ramp slope is:

$$V_o = \frac{10\ volts}{0.001\ sec} \tag{3}$$

$$V_o = 0.10\ volts/second \tag{4}$$

Triangle generators

Figure 2A shows a simplified circuit model of a triangle waveform generator. This circuit consists of a Miller integrator as the ramp generator, and an SPDT switch (S1) that can select either positive ($+V_{ref}$) or negative ($-V_{ref}$) reference voltage sources.

For purposes of discussion, switch S1 is an electronic switch that's toggled back and forth between positions A and B by a square wave applied to the control terminal (CT). Assume an initial condition (see **Figure 2B**) at time t2, at which point V_o = $-V1$, and the input of the integrator is connected to $-V_{ref}$. At time t2 the square wave switch driver changes to the opposite state, so S1 toggles to connect $+V_{ref}$ to the integrator input. The ramp output rises linearly at a rate of $+V_{ref}/RC$ until the switch toggles again at time t3. At this point the ramp is under the influence of $-V_{ref}$, so it drops linearly from +V1 to −V1. The switch toggles back and forth between $-V_{ref}$ and $+V_{ref}$ continuously, so the output (V_o) ramps back and forth between −V1 and +V1.

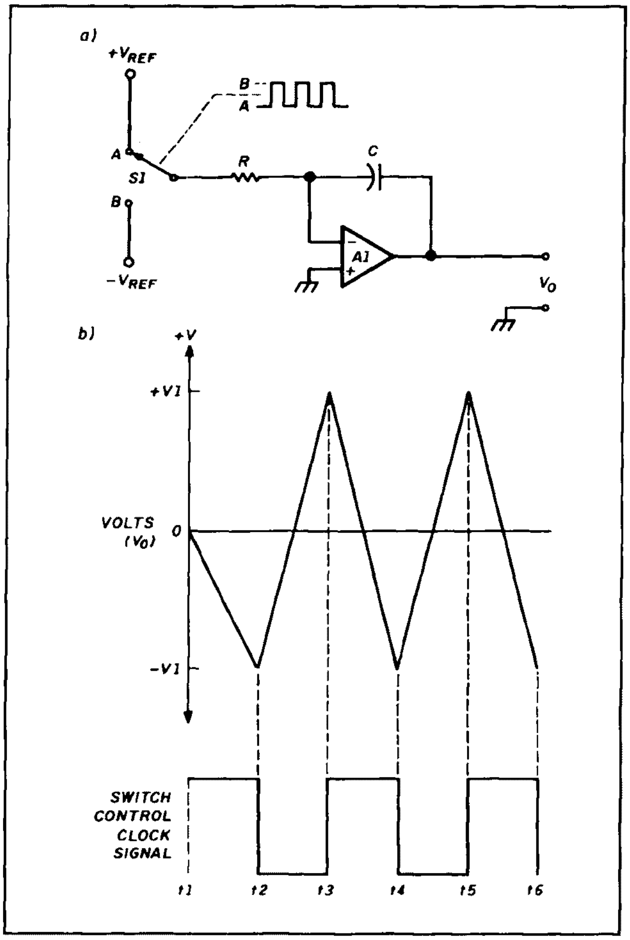
The circuit of **Figure 2A** isn't practical, but it serves as an analogy for the actual circuit. **Figure 3A** shows the circuit for a triangle waveform generator in which a Miller integrator forms the ramp generator and a voltage comparator serves as the switch. The comparator uses the positive feedback configuration, so it operates as a noninverting Schmitt trigger. Such a circuit snaps **HIGH** (V_B = $+V_{sat}$) when the input signal crosses a certain threshold voltage in the positive going direction. It will snap **LOW** again (V_B = $-V_{sat}$) when the input signal crosses a second threshold in a negative going direction. The two thresholds aren't always the same potential.

Because zener diodes CR1 and CR2 are in the circuit, the maximum allowable value of $+V_B$ is $[V_{CR1} + 0.7]$ volts, while the limit for $-V_B$ is $-[V_{CR2} + 0.7]$ volts. If V_{CR1} = V_{CR2}, then $|+V_B| = |-V_B|$. These potentials represent $\pm V_{ref}$ discussed in the analogy, so are the potentials that affect the ramp generator input.

Consider an initial state in which V_B is at the negative limit $-V_B$. The output V_o begins to ramp upwards from a minimum voltage of:

$$V1 = \frac{V_A(R2 + R4)}{R4} - \frac{V_B R2}{R4} \tag{5}$$

FIGURE 2



(A) Example of a Miller integrator used to create the constant current ramp generator. (B) The square wave signal going to the circuit of *A* and its resulting triangle wave output.

The output will continue to ramp upwards towards a maximum value of:

$$V3 = \frac{V2_A(R2 + R4)}{R4} + \frac{V_B R2}{R4} \tag{6}$$

causing a peak swing voltage of:

$$V_p = V3 - V1 \tag{7}$$

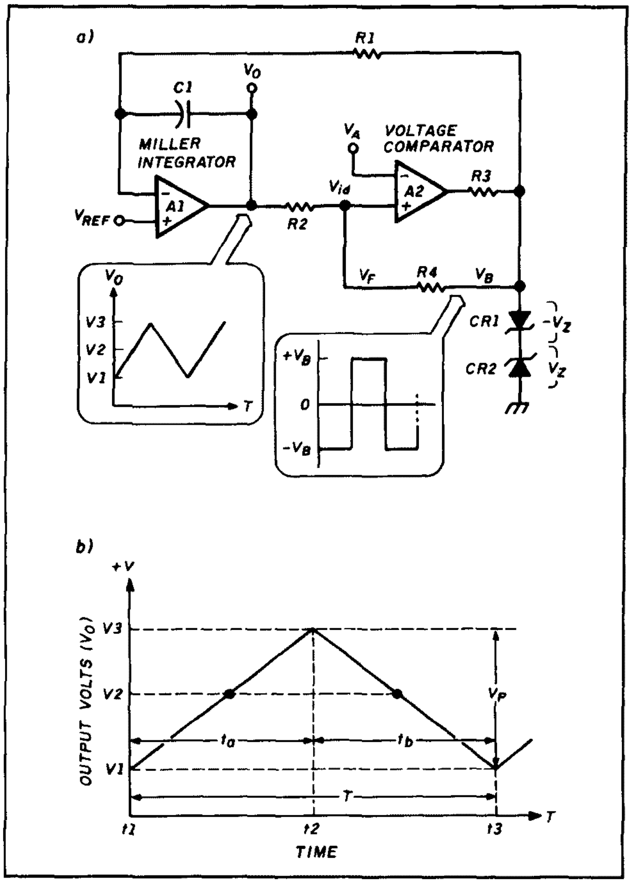
$$V_p = \left[\frac{V_A(R2 + R4)}{R4} + \frac{V_B R2}{R4}\right] - \left[\frac{V_A(R2 + R4)}{R4} - \frac{V_B R2}{R4}\right] \tag{8}$$

$$V_p = \frac{V_B R2}{R4} + \frac{V_B R2}{R4} = \frac{2V_V R2}{R4} \tag{9}$$

(See **Figure 3B** for details.)

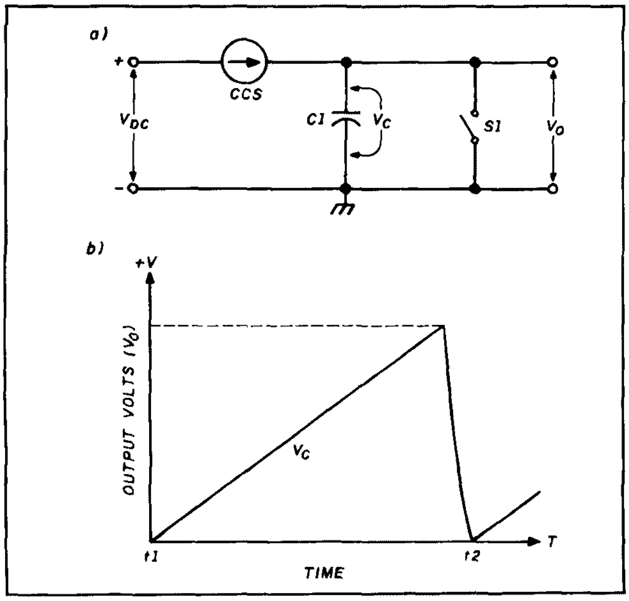
Comparator switching occurs when the differential input voltage V_{id} is zero. The inverting input (−IN) voltage is V_A,

FIGURE 3



(A) A practical circuit that produces the waveform seen in *Figure 2B*. (B) Detailed breakdown of one cycle of the triangle wave generator.

FIGURE 4



(A) Simple design of a sawtooth wave generator. (B) Output of the sawtooth generator; a CCS is used to charge the capacitor.

which is a fixed reference potential. The noninverting input (+IN) is at a voltage (V_F) that is the superimposition of two voltages, V_o and V_B:

$$V_F = \frac{V_oR4}{R2 + R4} + \frac{+/- V_BR2}{R2 + R3} \qquad \textbf{(10)}$$

If $+V_B = -V_B$, then the positive and negative thresholds are equal.

The duration of each ramp (t_a and t_b) can be found from:

$$t_{a,b} = \frac{V_p}{\left| \dfrac{V_B}{R1C1} \right|} \qquad \textbf{(11)}$$

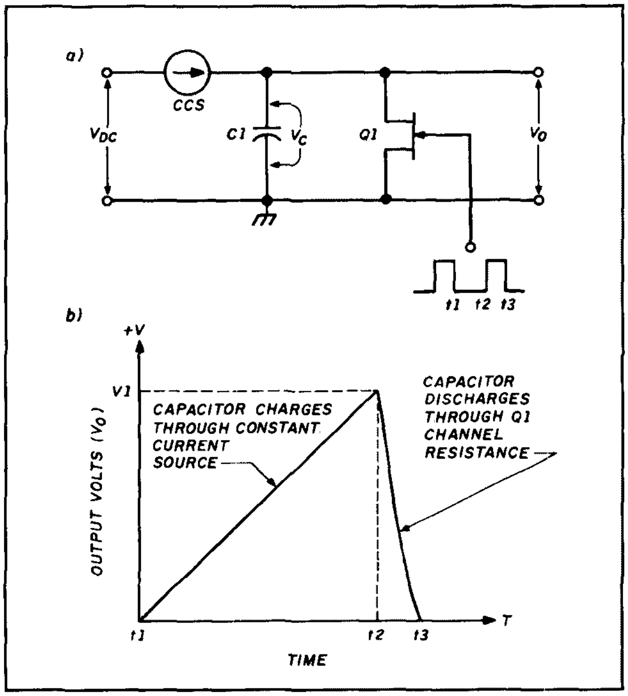
The value of V_B is selected from $-V_B$ or $+V_B$ as needed. In Equation 10 you find that $V_p = 2V_BR2/R4$, so:

$$t_{a,b} = \frac{\left| \dfrac{2V_BR2}{R4} \right|}{\left| \dfrac{V_B}{R1C1} \right|} \qquad \textbf{(12)}$$

$$t_{a,b} = \left| \frac{R1C1}{V_B} \right| \left| \frac{2V_BR2}{R4} \right| \qquad \textbf{(13)}$$

$$t_{a,b} = R1C1 \left| \frac{2V_BR2}{R4} \right| \qquad \textbf{(14)}$$

FIGURE 5



(A) An example of a periodic sawtooth oscillator. (B) The same basic wave output.

or, in the less general (but more common) case of $t_a = t_b$:

$$T = 2R1C1\left(\frac{2R2}{R4}\right) \qquad \textbf{(15)}$$

The frequency of the triangle wave is the reciprocal of the period (1/T), so:

$$F = \frac{1}{T} \qquad \textbf{(16)}$$

$$F = \frac{1}{\frac{4R1C1R2}{R4}} \qquad \textbf{(17)}$$

$$F = \frac{R4}{4R1C1R2} \qquad \textbf{(18)}$$

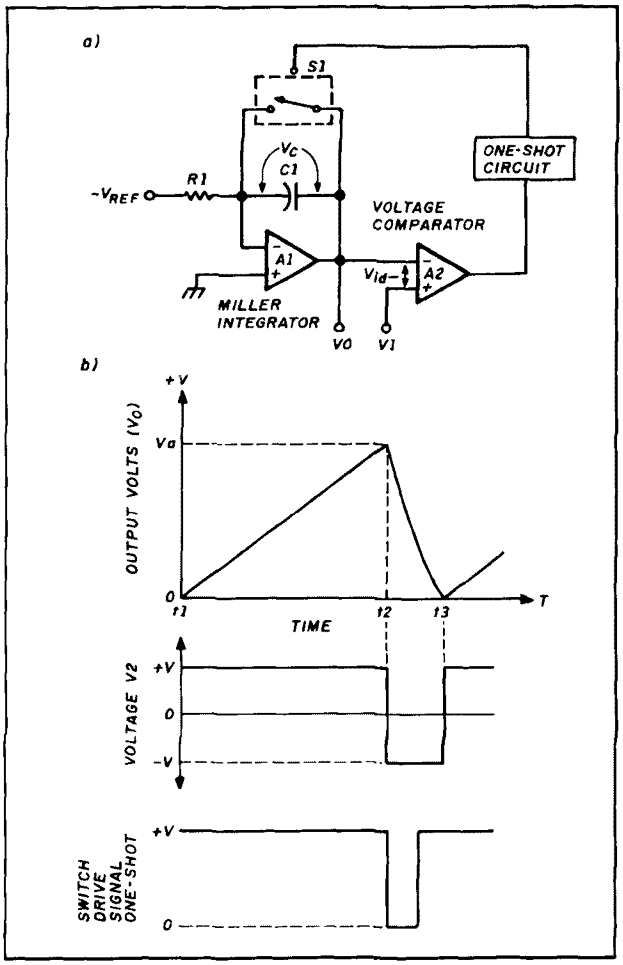
Sawtooth generators

The *sawtooth wave* in **Figure 1A** is a single slope ramp function. The wave ramps linearly upwards (or downwards), and then snaps back abruptly to the initial baseline condition. **Figure 4A** shows a simple model of a sawtooth generator circuit. A constant current source charges a capacitor in a manner that generates the linear ramp function (see **Figure 4B**). When the ramp voltage (V_c) reaches the maximum point (V_p) switch S1 is closed, forcing V_c back to zero by discharging the capacitor. If switch S1 remains closed, the sawtooth is terminated; if S1 reopens, a second sawtooth is created as the capacitor recharges.

Figure 5A shows the circuit for a periodic sawtooth oscillator. It's similar to **Figure 4A** except that a junction field effect transistor (JFET), Q1, is used as the discharge switch. When Q1 is turned off, the output voltage ramps upwards (see **Figure 5B**). When the gate is pulsed hard on, the drain source channel resistance drops from a very high value to a very low one, forcing C1 to discharge rapidly. In the absence of a gate pulse, however, the channel resistance remains very high. The gate is turned off at time t1, so V_c begins ramping upwards. At t2 the JFET gate is pulsed, so C1 rapidly discharges back to zero. When the pulse (t2-t3) ends, Q1 turns off again and the ramp starts over. You can use the same circuit for single sweep operation by replacing the pulse train applied to the gate of Q1 with the output of a monostable multivibrator.

The circuit of **Figure 6A** shows a sawtooth generator that uses a Miller integrator (A1) as a ramp generator and replaces the discharge switch with an electronic switch driven by a voltage comparator and one-shot circuit. The timing diagram for this circuit is shown in **Figure 6B**. Under the initial conditions, at time t1, the output voltage (V_o) ramps upwards at a rate of $[-(-V_{ref})/R1C1]$. The voltage *comparator* (A2) is biased with the noninverting input (+IN) set to V1 and the inverting input at V_o. The comparator differential input voltage $V_{id} = (V1 - V_o)$. As long as $V1 > V_o$, the comparator sees a negative input and produces a **HIGH** output of $+V_{sat}$. At the point where $V1 = V_o$, the differential input voltage is zero, so the output of A2 (voltage V2) drops **LOW** (i.e., $-V_{sat}$). The negative going edge of V2 at time t2 triggers the one-shot circuit. The output of the one-shot closes electronic switch S1 briefly, causing the capacitor to discharge. The one-shot pulse ends at time t3, so S1 reopens and allows V_o to ramp upwards again.

FIGURE 6

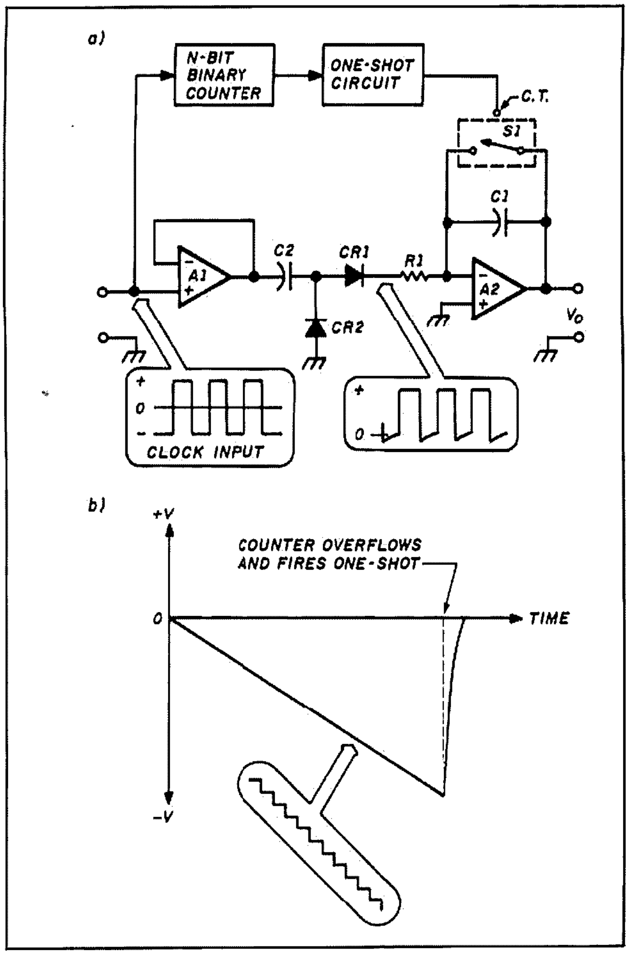


(A) A Miller integrator used as a sawtooth generator. The discharge switch is replaced by an electronic switch driven by a voltage comparator/one-shot circuit. (B) Timing diagram for the circuit in *Figure 6A*. At time t2, the negative going edge of V2 triggers the one shot.

The *staircase generator* in **Figure 7A** is a variant of the sawtooth generator circuit. The input amplifier (A1) provides buffering. A square wave clock signal applied to the input of A1 is passed through capacitor C2 to a diode-clipping network (CR1, CR2). The clipping circuit removes the negative excursions of the square wave (see inset to **Figure 7A**). The remaining positive polarity pulses are applied to the input of the inverting Miller integrator ramp generator circuit. Each pulse adds a slight step increase to the capacitor charge voltage, so (unless there is significant droop between pulses) the output will ramp up to a negative potential in the staircase fashion shown in the inset to **Figure 7B**.

The reset circuitry in this circuit is a little different. Although the comparator method of **Figure 6A** would also work, this circuit takes advantage of the input square wave to provide the period timing of the sawtooth. The square waves are

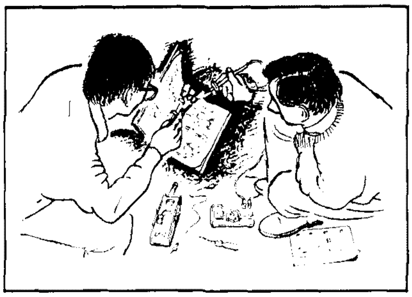
FIGURE 7



(A) Circuit of a staircase generator. (B) Example of the staircase generator output.

applied to the input of an N-bit binary digital counter circuit. When 2^N pulses have passed, the counter overflows on $[2^N+1]$ and triggers a one-shot circuit. As in the previous case, the one-shot output pulse closes the electronic reset switch shunted across capacitor C1 momentarily.

Ham Notebook



The 80/40-meter Junkbox Rig Revisited

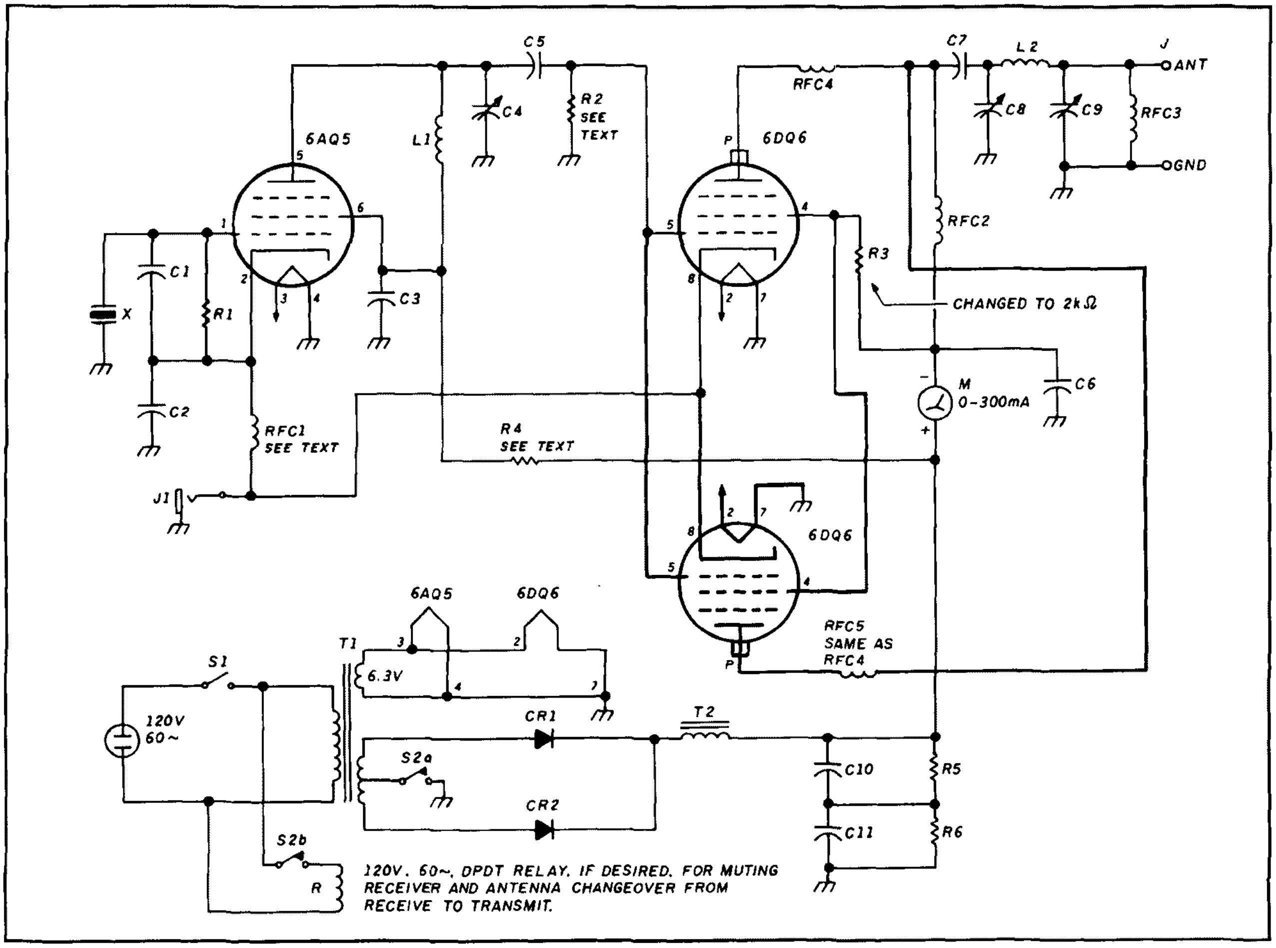
Some of you may have built the transmitter I described in the December 1989 issue of *Ham Radio*.[1] I hope you've enjoyed this little rig, but perhaps you wish you had just a little more power. Fret no more. This may be just what the doctor ordered — *parallel* 6DQ6s.

All you need to do is add another final amplifier tube as shown in the schematic diagram in **Figure 1**. If your milliammeter covers only 0 to 150 mA, the range should be doubled to 0 to 300 mA. In other words, if you now load your rig to 125 mA, you should be able to load it to 250 mA. It's possible to almost double your power, or increase it by 3 dB.

Parts to be changed

You must change the following parts. If you refer to the original diagram,

FIGURE 1



Upgraded 80/40-meter Junkbox Rig.

you'll find that R2 is a 27-k resistor, R3 is a 4-k wire-wound resistor, and R4 is a 1.5-k wire-wound resistor. These should be halved. You can reduce R2 to about 12 to 15 k. However, I tried this and the change seemed to make no difference. R3 should now be 2 k; R4 could be either 500 to 1000 ohms, or simply deleted.

I don't use a voltage-dropping resistor for R4 now as I don't exceed 300 volts with this parallel arrangement. You should change RFC1 to 300 mA or move the 6DQ6 cathode wire from pin 8 of the octal socket directly to the ungrounded side of the key jack.

New currents and voltages

The following currents and voltages result from this change:

• 200 mA × 285 volts = 57 watts input. (Formerly 100 mA × 305 volts = 32 watts input.)

• 250 mA × 280 volts = 70 watts input. (Formerly 125 mA × 300 volts = 38 watts input.)

• 300 mA × 275 volts = 83 watts input. (Formerly 150 mA × 295 volts = 44 watts input.)

Loading is the same as before. If you follow the directions in December's article, you shouldn't encounter any difficulties.

Charlie Tiemeyer, W3RMD

REFERENCE

1. Charlie Tiemeyer, W3RMD, "The Five-Band Junkbox Transmitter," *Ham Radio*, December 1989, page 42.

Ezy tune

While tuning around the HF bands one night with my Heath HW-5400, I discovered that I needed an easier way to change frequency than the conventional tuning knob. I came up with a little circuit I call the "Ezy Tune" (see **Figure** 1). It's easy to build and doesn't affect the manual tuning capabilities of the HW-5400.

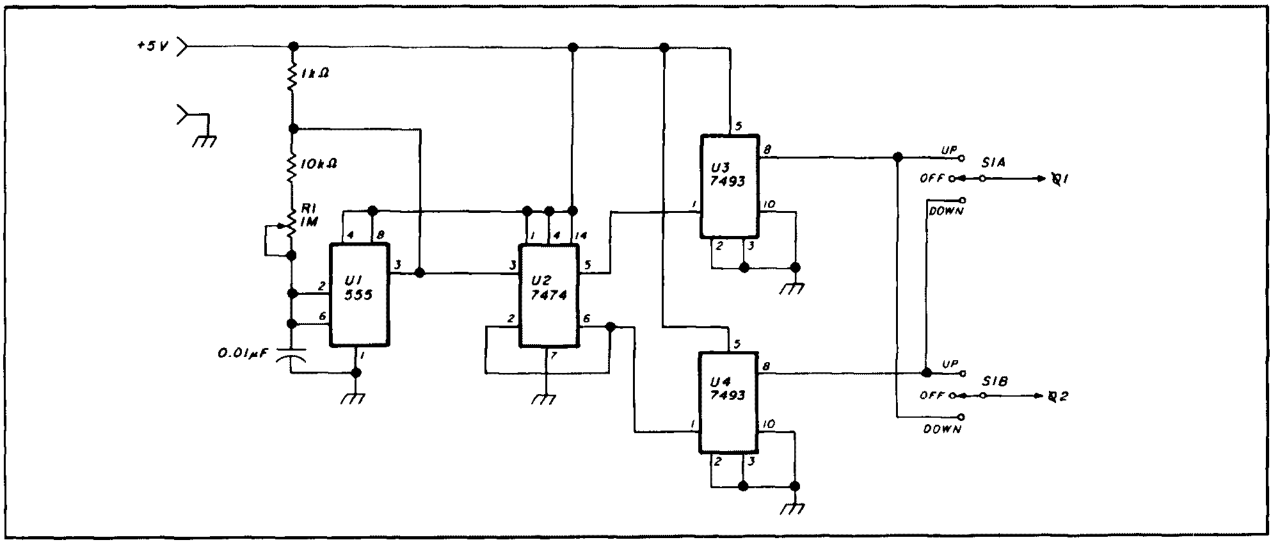
The master oscillator, U1, is a 555 astable multivibrator free running at eight times the tuning rate in steps per second, as desired. The frequency of U1 is adjustable by R1, a 1-meg pot. U2 produces two pulse trains 180 degrees out of phase which, when divided by the two binary counters U3 and U4, deliver two pulse trains 90 degrees out of phase. This is what the optical encoder controlled by the tuning knob on the HW-5400 does. These two signals are switched through S1, a DPDT center-off toggle switch, to the θ1 and θ2 inputs on the controller board.

I built the Ezy Tune circuit on a piece of perfboard small enough to mount near the controller board. Power can be supplied by P703 on the controller board; pin 1 is 5 volts and pin 3 is common. S1 can be panel mounted to the lower left or right of the tuning knob.

By the way, this circuit would probably work with other transceivers with optically encoded tuning systems using TTL-level inputs.

Dexter King, AB4DP

FIGURE 1



The schematic of the "Ezy Tune" circuit.

THE QRP TLC-KEYER

By Rick Littlefield, K1BQT, Box 114, Barrington, New Hampshire 03825

Preparing for a QRP DXpedition usually means packing to travel light. Unfortunately, adding even a few basic accessories can overload the radio bag in a hurry. Take electronic keyers, for example. Mine is larger than my QRP transceiver, and loaded with features I rarely use. To make life on the road less awkward, I decided to build a miniature keyer — one designed specifically for portable use. Not wanting to spend a lot of money, I settled on a single-paddle type employing inexpensive ICs available from Radio Shack.

The circuit

The design I chose is a popular one using NE555 timers.[1] The circuit performed well on the breadboard, but required over 20 mA at 9 volts to operate. This was too much current for sustained operation with a small 9-volt battery. To solve this problem, I substituted TLC555s — a CMOS direct replacement for the NE555 (hence the name TLC-Keyer). This change reduced continuous current drain to a scant 1.4 mA, making battery operation practical and allowing me to include a reed relay output circuit. I prefer using reed switching because it interfaces with virtually any rig.

The circuit consists of three simple timer stages. U1 generates the space interval between characters, while U2 and U3 generate dot and dash outputs, respectively. A 10-k linear pot varies sending speed by raising or lowering the timing interval of all three chips simultaneously. Spacing and character timing may be altered by substituting new values for the 33 and 82-k series resistors in this portion of the circuit. For example, changing the 82-k resistor to 100 k results in longer dashes.

A 2N2222 DC switch actuates relay K1 whenever the output of U2 or U3 goes high. For self-powered operation from a 9-volt battery, a 470-ohm current limiting resistor is connected in series with the 5-volt relay coil. For 12-volt operation, use 1.2 k or install a jumper and a 12-volt relay.

Construction

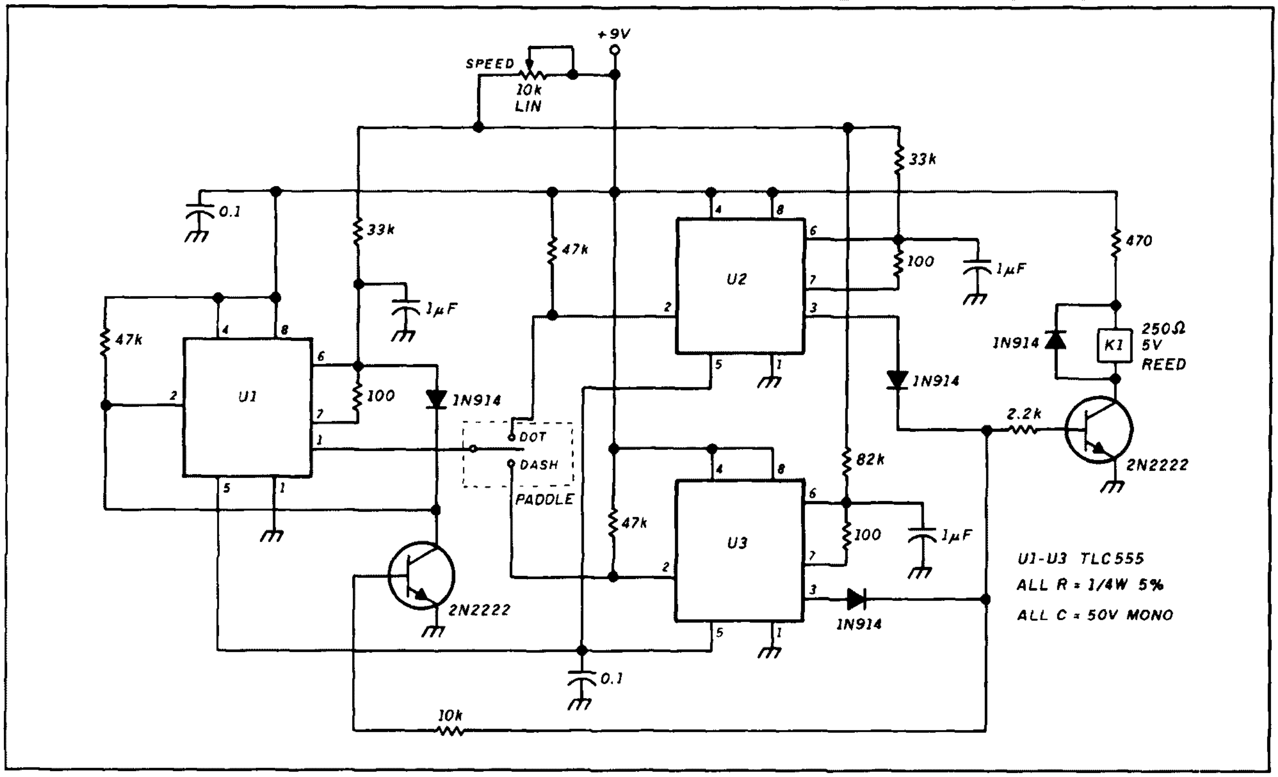
Board construction is straightforward — simply follow the parts placement diagram shown in **Figures 1A** through **1C**. Remember to observe CMOS handling precautions when installing the TLC555s. Also, install jumpers and interconnecting leads.

Details for constructing and mounting the paddle assembly are shown in **Figure 2**. I fabricated the paddle arm from scrap G-10 board (all copper removed). The key contact is a 1/4-inch no. 4-40 screw and no. 4 nut. I filed the screw tip flush with the nut, and polished it to make a better contact surface. Lexan™ or Styrene™ sheet stock makes a good paddle, and is readily tapped to accept no. 2-56 or no. 4-40 mounting screws. Standard 3/8-inch no. 6 spacers and 90-degree L brackets support the paddle assembly above the pc board components. Two 3/4-inch no. 4 spacers serve as dot and dash contacts. Contacts should be roughly aligned and screws tightened before the pc board is installed in a cabinet or on a base.

Final packaging depends on how you intend to power the keyer. One option is to mount the board, paddle, and battery in a small project box. This lets you operate on internal power and use the keyer with any rig. The second option is to borrow power from your transceiver — replacing the key jack with a stereo jack, and using the tip connection to supply 12 volts. Obviously, eliminating the battery will reduce the size and weight of the keyer package.

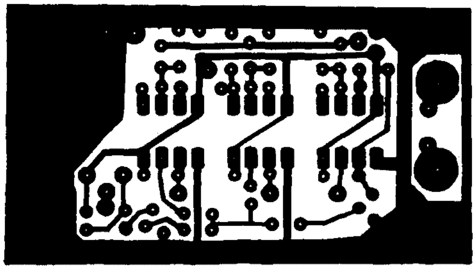
I don't recommend mounting the keyer electronics inside your rig. For one thing, the "common" key line of this circuit is the output of U1, and this must remain isolated from ground. Also, the pc board fits neatly on the keyer base without contributing to its size. No real space saving is gained from putting it inside the radio. Whether you package the keyer in a project box or on a weighted base plate, cement a rubber pad to the bottom. This will reduce skidding on smooth surfaces.

FIGURE 1A



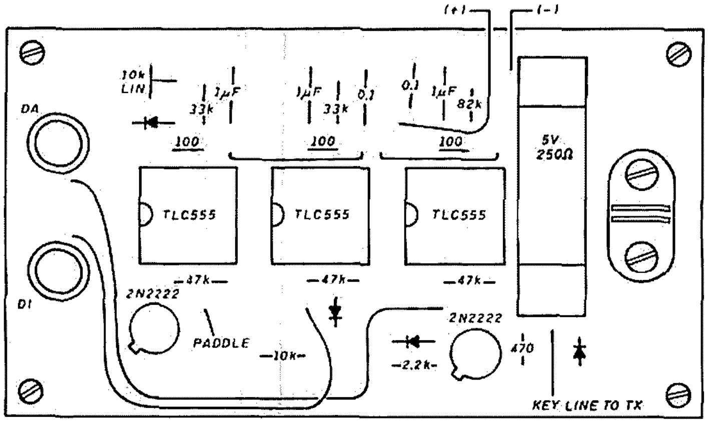
The circuit schematic.

FIGURE 1B



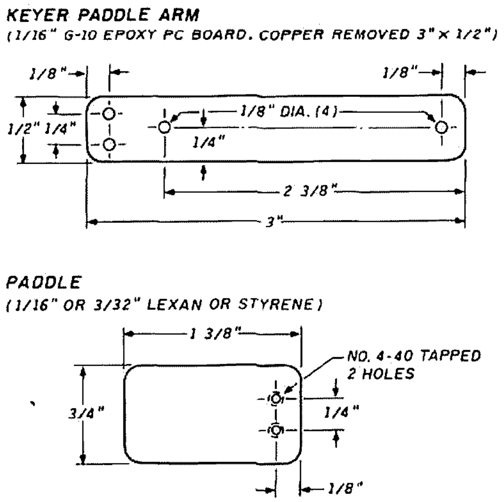
The foil side of the pc board.

FIGURE 1C



The component placement guide.

FIGURE 2



The paddle assembly.

PARTS LIST

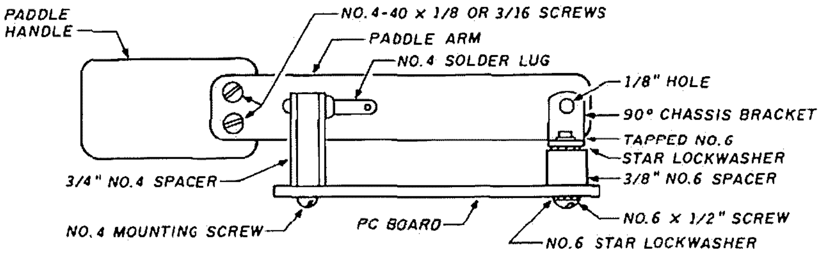
Capacitors
2 0.1 µF Mono
3 1 µF Mono or equivalent

Resistors
3 100
1 2.2 k
1 10 k
2 33 k
3 47 k
1 82 k
1 10-k linear pot

Semiconductors
3 TLC555
2 2N2222
4 1N914

Miscellaneous
1 pc board (Available from Far Circuits, 18N640 Field Court, Dundee, IL 60118, for $4.00. Price includes shipping and handling.)

FIGURE 3



Dimensions of the completed paddle assembly.

Operation

Keyer setup is purely mechanical. To make fine centering adjustments on the arm, place the tip of a flat-blade screwdriver between the two support spacers and torque the arm gently in the desired direction. When you've completed this operation, you may want to readjust the dot/dash posts for the desired contact spacing. Finally, connect the keyer output to an oscillator or rig, set the speed control midway, and start sending. If the keyer fails to operate, confirm that you have power and that diodes, chips, and transistors are positioned correctly.

Conclusion

The finished key, shown in Figure 3, measures 2.6 by 1.5 by 1.5 inches, and weighs only a few ounces. This represents a vast improvement in portability over my old one! The keyer works well — with one minor reservation. I had used the Curtis 8044s for years, and the TLC's unbuffered input threw me on my first sending attempts. With no buffer, the keyer can't accept a dot while a dash is completing. Consequently, I rapped out a few "CQ DE M1BOTs" before I brought my thumb under control. However, I adapted quickly and can now switch from one keyer to the other without difficulty.

The TLC-keyer is now a permanent part of my QRP station, and I travel several pounds lighter for my effort. If you face a similar dilemma at packing time, give it a try. The price is right, and so is the size! hr

REFERENCES

1 Wes Hayward, W7ZOI, and Doug DeMaw, W1FB, *Solid-State Design For the Radio Amateur*, ARRL, 1985, pages 177-178.

(continued from page 9)

was also done in the familiar "brunswick green," but it boasted more selectivity and punch. It was the old S-40A. I saved my money for it a long time and finally, when it seemed I could scrape together no more, my parents matched my funds so that I could have it for a birthday gift.

The order was placed with Bob Henry out in Butler, Missouri and the radio was promised for delivery in time for my October birthday. In those days there was no UPS, and a radio the size of the S-40A had to be shipped by Railway Express. Fate intervened in the form of a nationwide railroad strike, and it appeared my gift would arrive long after the birthday had passed.

During those weeks of waiting, my father called Bob Henry several times. When he learned the special occasion for which the radio had been ordered, Bob went the second mile. He put a tracer out with Railway Express and located the radio. It was well over two-thirds of its journey to my South Carolina home, and Bob assured my father that it should reach our town only a day or two after the strike was settled. About that time President Truman intervened, the strike was halted, the trains rolled again, and the radio arrived in Sumter, South Carolina on the evening train from Atlanta at 9:05 p.m. on the evening of my 16th birthday!

That radio served me well, even after I acquired my ham ticket several years later. I eventually used it as trade-in towards a better model, the SX-42, but it always held a special place in my heart — even though it was not my first receiver.

The products Bill Halligan's Chicago factory turned out will, I expect, always be held in special esteem by those of us old enough to remember them. I later owned a couple of Hammarlund receivers, which were also very fine radios. But somehow, there was a special mystique about the Hallicrafters that even the National HRO couldn't touch.

Even though Bill's gear no longer graces our shacks, Bob Henry has joined the ranks of the Silent Keys, and the Butler store is gone, there is one ham who will never forget the touch both of them had on his life.

Drayton Cooper, N4LBJ,
Bowling Green, South Carolina

short circuits

Missing Parenthesis

In WA3EKL's October 1989 article "The PV-4 On Your Commodore," there was a small error in the proof program on page 64. Line 170 should be corrected to read:

```
F1=(1-((10.7575*(LOG(KA)/LOG
(10)))-8) Λ -1)/2*A(j))
```

Program Correction

In N1AYW's article, "Computerizing Smith Chart Network Analysis" (October 1989, page 10), Menu item 6 in the resonant circuit calculations of the Smith Chart program is not functional as it was originally listed; however, the error was corrected in version 2.46. For those readers with versions earlier than 2.46, please insert the following two lines and modify line 3400 as shown below.

```
2915 FR=10 Λ 6/(2*PI*CR*RR*QR)
:RETURN:REM SERIES, C KNOWN
2925 FR=QR*10 Λ 6/(2*PI*RR*CR)
:RETURN:REM PARALLEL,
C KNOWN
3400 GOSUB 2830:GOSUB
2850:GOSUB 2790:ON PC GOSUB
2915,2925:GOSUB 2930:PRINT"F
="FR" and L ="LR:GOSUB
3010:GOTO 2870
```

Short circuit for Tseng Liao

In the Ham Note, "Two in one: Trace doubler for CRO, and square wave and pulse generator," published in the December 1988 issue, there are two misprints:

(a) In Figure 3A, C_4 should be C_p and R_{13} should be R_p.

(b) In the "Practical examples" section, R_1 should be R_i.

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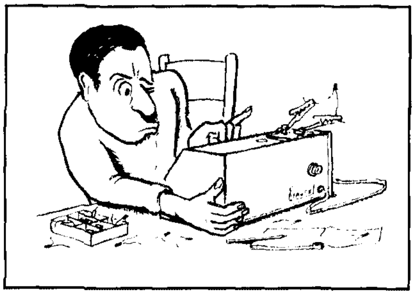
Practically Speaking

Joseph J. Carr, K4IPV

GETTING TO KNOW THE LOGIC FAMILIES — PART 2: CMOS



Complementary metal oxide semiconductor (CMOS) digital ICs are well known and readily available. Designated by "4xxx" series-type numbers, CMOS devices offer extremely low current drain, moderate operating speeds, and low cost. The low current requirements (on the order of a few microamperes) of CMOS technology have made possible a wide range of portable products — from digital watches to calculators to laptop microcomputers. Because of their properties, CMOS devices also make very useful sensing circuits for a variety of applications not always associated with digital electronics. But before getting into the various circuits, let's review the basics of CMOS digital IC devices.

CMOS digital devices

The MOSFET transistor is the basis of the CMOS line of digital ICs. This device offers an extremely high input impedance because the control gate input is isolated from the "channel" by a layer of insulation (see **Figure 1A**).

There are two polarities of MOSFET transistor, determined by the type of semiconductor material used in the channel structure. If N-type material is used, the MOSFET is an N-channel; if P-type semiconductor material is used, it's a P-channel MOSFET. The respective circuit symbols for these two devices are shown in **Figure 1B**.

In digital circuits, the MOSFET will be in one of two conditions. When a bias is applied to the gate, the channel resistance is very high (megohms). On the other hand, the absence of bias causes the channel resistance to be very low (200 to 300 k). Although this explanation doesn't include all possible types and is somewhat simplified, it will do for our purposes.

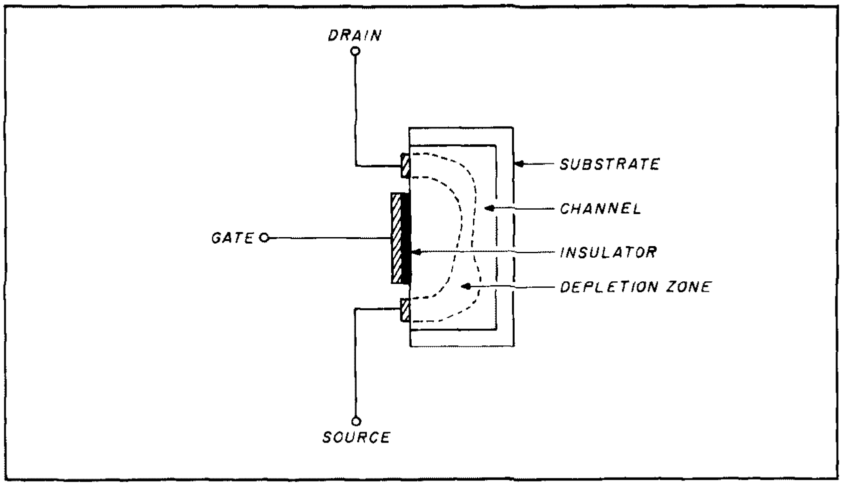
When examining any digital logic element it's useful to consider the case of the inverter as representative of the whole class. **Figure 2** shows the circuit of a CMOS inverter. This circuit consists of an N-type MOSFET and a P-type MOSFET connected with their respective gates in parallel and their channels in series. The input of the inverter is the gate; the output is the junction between the P-channel drain (Q2) and the N-channel source (Q1).

Two DC power supplies are shown: V+ and V−. These voltages are typically ±4.5 to ±15 volts DC. In some cases, the V− supply is set to zero, so the V− terminal of the device will be simply grounded.

Because P and N-channel MOSFETs are of opposite polarity, one will have a high channel resistance while the other has a low resistance. The two channels are connected in series, so the overall channel resistance (as measured from (V+) terminal to (V−) terminal) is very high. This resistance is why the CMOS device draws such low current — the power supply sees a resistance in the megohm range. The only time the CMOS devices draw appreciable current is during output transition from high-to-low or low-to-high. At that brief instant, both channel resistances are in a transition region between high and low resistance values.

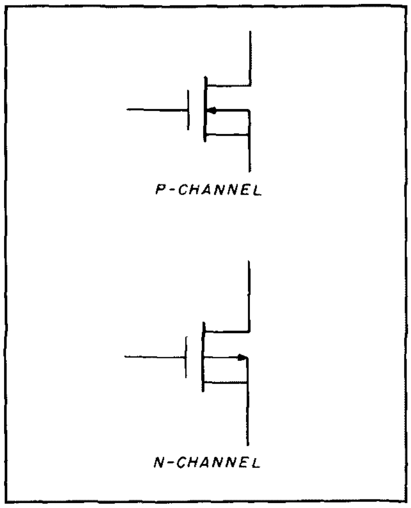
Figure 3 is a graphical representation of the channel resistance relationship.

FIGURE 1A



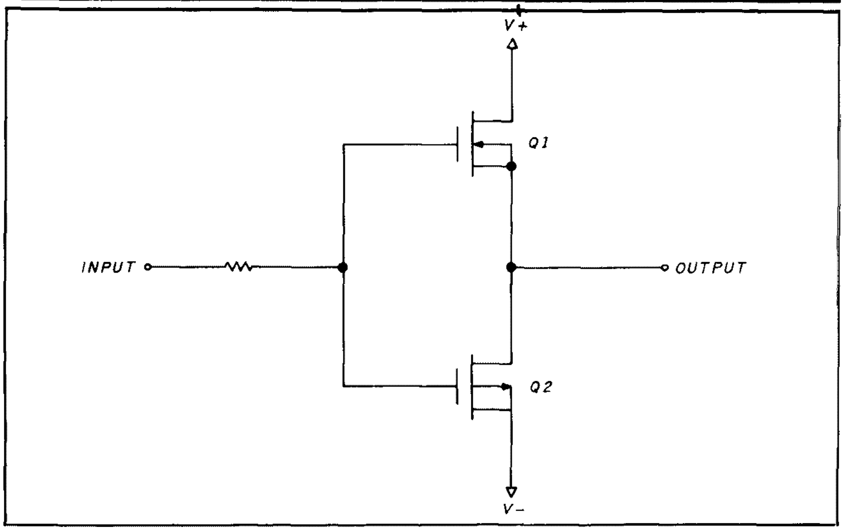
Physical makeup of a MOSFET transistor. These transistors comprise most CMOS devices.

FIGURE 1B



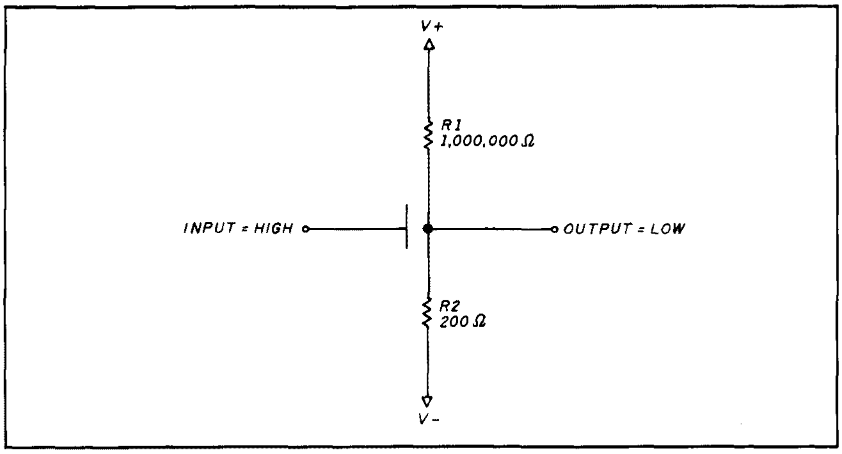
P-channel and N-channel MOSFET transistor designators.

FIGURE 2



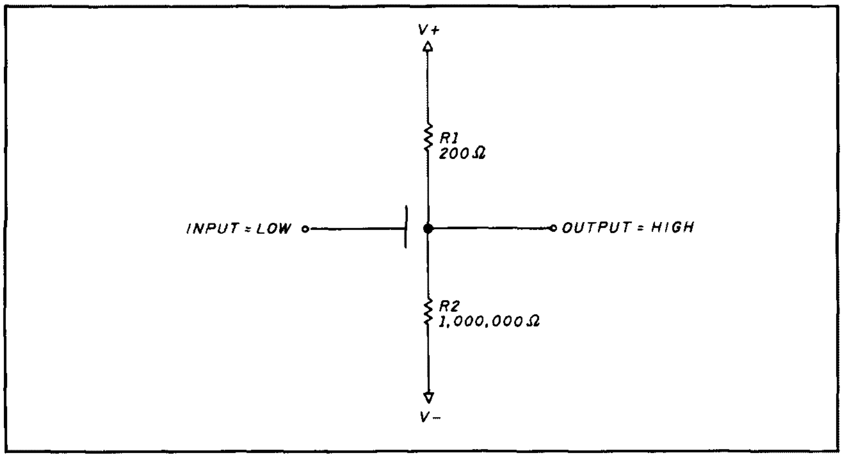
Schematic of a typical CMOS inverter.

FIGURE 3A



Situation where the input of the CMOS inverter is HIGH.

FIGURE 3B



Situation where the input of the CMOS inverter is LOW.

In the cases shown in **Figures 3A** and **3B**, the total resistance (R1 + R2) is the same even though the relationship of R1 and R2 changes. In **Figure 3A** the input of the CMOS inverter (**Figure 2**) is high. The resistance of Q1 (i.e., R1) is high and that of Q2 is low. Thus, the output line is connected through a low resistance to V−. In **Figure 3B** the input is low. Here, the relationship of R1 and R2 is reversed. R2 is high, while R1 is low. This means that the output terminal is connected to the V+ supply through a low resistance.

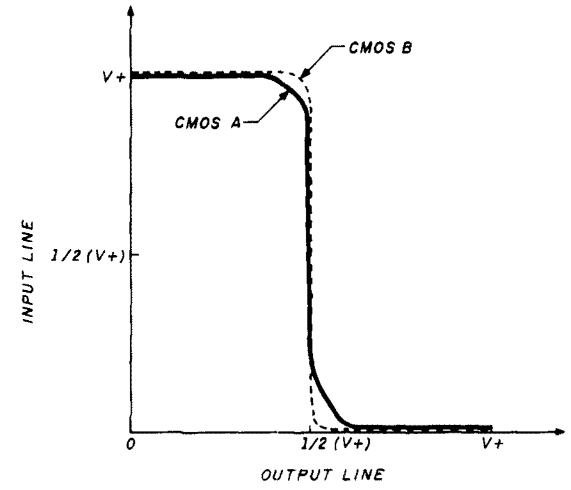
The definitions of logical high and logical low states are determined by the voltages used. If two potentials are used, V− is low and V+ is high. But if the V− is set to zero, logical low is zero (grounded), while the logical high remains at V+. The transition point between high and low (or vice versa) occurs when the input is at a potential halfway between V+ and V−, or when V− is zero volts (one-half V+).

Electrostatic discharge (ESD) damage

The insulated gate of a MOSFET transistor is very thin. The ability to withstand high voltages is directly proportional to the thickness (for any given material). This thin layer means MOSFETs can handle only 50 to 150 volts (80 volts is common). Greater potentials will pop through the gate insulation, destroying the device. Ordinary handling of tools and other implements can create potentials up to several kilovolts. These potentials can build up in the human body, too. If you don't believe me, walk across a carpet on a dry winter day and then touch a grounded object!

Unfortunately, the damage from ESD doesn't always happen immediately. It can show up as an unexplained "spontaneous" failure sometime later. The usual procedure to minimize ESD damage is to make sure that all pins are at the same potential all of the time. This usually means working on a metal or carbonized foam surface, and storing the devices in a conductive container or on carbonized foam pads. It's also recommended that you discharge the potentials on your body by touching something grounded (but not in the presence of high voltage or AC) just before you touch the MOSFET or CMOS device. Actually, it's a good idea to avoid handling the device at all unless absolutely necessary.

FIGURE 4



Transition time differences between A-series and B-series CMOS devices.

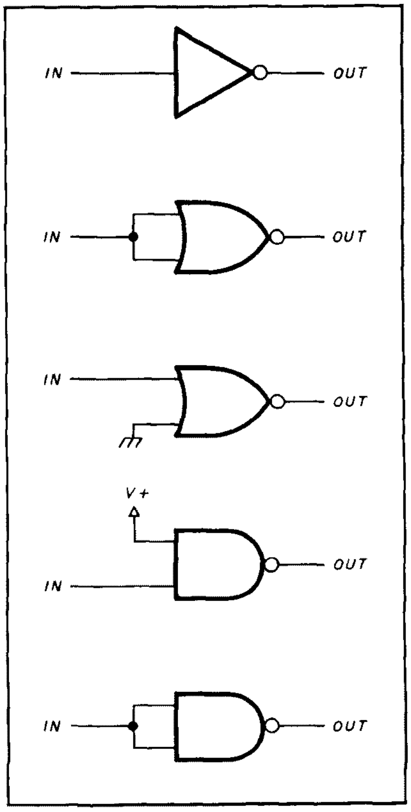
A-series versus B-series CMOS

There are two different types of CMOS devices on the market: A-series and B-series chips. The A-series is the older of the two. These devices are designated either with an "A" suffix on the type number, or no suffix on the type number. For example, both "4013" and "4013A" are designators for an A-series type 4013 device. The B-series is newer; some believe it's an improvement over the A-series. B-series devices are always marked with a "B" suffix on the type number (e.g., 4013B).

There are several notable differences between A-series and B-series devices. Perhaps the most well known is that B-series devices are protected from ESD damage somewhat by internal zener diodes which clamp or bypass the high voltage electrostatic potentials. You can get away with handling B-series devices more often than A-series. Note that the amount of ESD damage on B-series does not drop to zero, but it's very low compared with A-series.

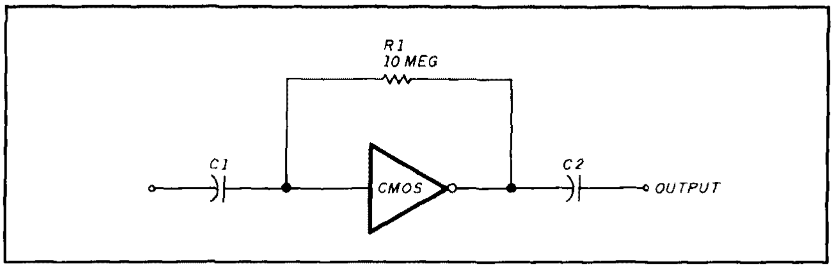
Another difference is the output transition time. Figure 4 shows the transfer function of the two types of CMOS devices. The B-series makes the transition more quickly, creating a sharper, faster rise time pulse. A final difference

FIGURE 5



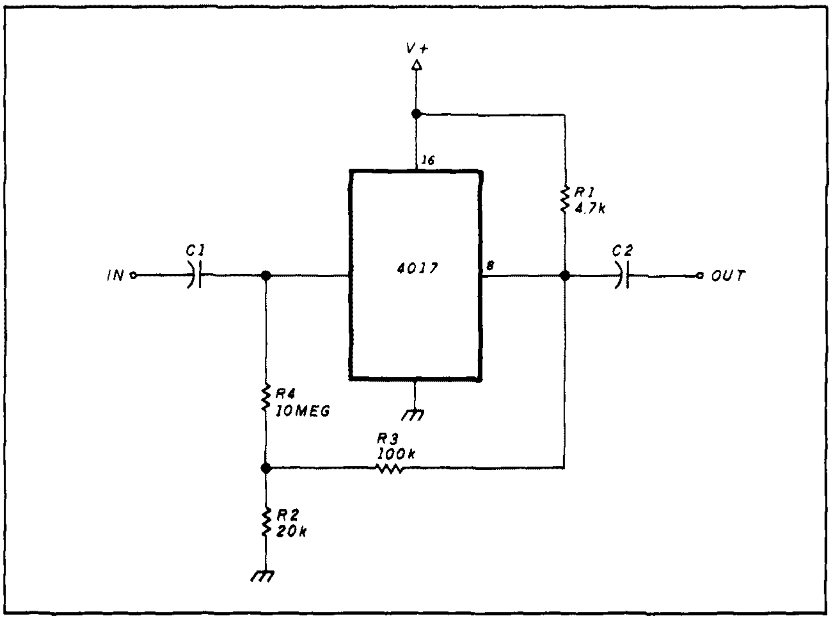
Several methods of "making" an inverter from CMOS NOR and NAND gates.

FIGURE 6A



An elementary biasing scheme for using a CMOS device in the linear mode.

FIGURE 6B



A 4017 CMOS device used as an audio amplifier.

is that most B-series CMOS devices will drive a larger load than their A-series cousins. In one case, for example, the B-series will drive a load three times heavier than the equivalent A-series device.

CMOS inverters

Most of the circuits in this article are based on the inverter. There are several inverters in the CMOS line. You can also make an inverter from available NOR or NAND gates. **Figure 5** shows how to make an inverter from other logic elements. For both the NOR and NAND gates, connecting the inputs together will cause the element to work. You could also make an inverter by grounding one input of a NOR gate, or connecting one input of a NAND gate high. Another alternative is to use a CMOS transistor array like the CD-3600 or the 4017 device.

Linear operation of CMOS devices

You can make the CMOS digital IC into a linear device using appropriate biasing methods. **Figure 6** shows two examples of linear operation of a CMOS inverter.

The elementary bias method is shown in **Figure 6A**. This circuit is quite simple. It consists of a pair of coupling capacitors (one on the input and one on the output) and a feedback resistor between output and input. It's possible to exert a small degree of control over linearity by varying the 10-meg value.

The circuit of **Figure 6B** is an audio amplifier based on the CMOS transistor array, type 4017. The 4017 array consists of three independent N-channnel/P-channel complementary pairs. In this project, I'm using one of the 4017 transistors. The bias is derived in a similar manner from the output, and also from the V+ power supply.

The values of input and output capacitance for both circuits depend on the lower −3 dB frequency response desired. This capacitance is set according to **Equation 1** if you use 10 megs for the resistance seen by the input signal (as in **Figure 6**):

$$C\mu F = \frac{0.1}{2\pi F} \qquad (1)$$

Where:

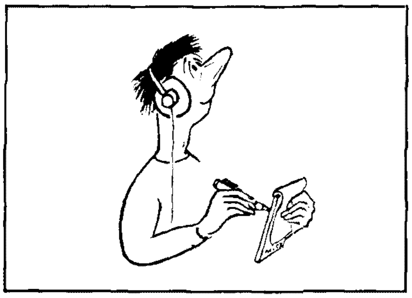
$C\mu F$ is the capacitance in microfarads

F is the −3 dB point in the desired frequency response curve, in hertz

Conclusion

Although CMOS devices are normally thought of as "digital," their unique properties make them useful in a lot of other applications. Understanding the properties of the CMOS device will help you make them work in other than normal circuits. hr

DX Forecaster



By Garth Stonehocker, KØRYW

THE DX BEACON GAME

Some radio stations can be used as beacons for DX propagation and research studies. There are three criteria for a radio station to be a beacon: its location, its identification (including frequency and modulator), and a useful schedule — in that order of importance. Most frequency ranges (VLF, LF, MF, HF, and VHF) contain radio stations which are used as beacons. In the LF range, the frequency band from 200 to 400 kHz is set aside for beacon stations used by marine and aeronautical radio direction-finding receivers for determining position location. The MF/AM broadcast stations form a secondary group of beacons also for this purpose. Other frequency ranges are generally used for other purposes and meet enough of the criteria to be useful as beacons only by coincidence. Of course, Amateur Radio, research, and government communities have set up specific radio stations as beacons in many locations on many frequencies with usable schedules.

The frequency ranges useful for ionospheric propagation are generally MF, HF, and the lower VHF. The beacons in these frequency ranges are mostly individual radio stations run by Amateurs, broadcasters, researchers, and governments. Of note is an Amateur network, PROPNET, with several locations using one frequency — 14.1 MHz. Each of ten stations transmits an identification in CW on a 1-minute time schedule with three power levels. The stations are located in New York, San Francisco, Hawaii, Japan, Israel, Finland, Madeira Island, South Africa, Buenos Aires, and Columbia. Not all of these stations are on the air consistently, so their minute time slot is not always in use.

The radio stations of world time and frequency standards offer other opportunities which can be used in the same way. The frequencies set aside for this purpose are generally 5, 10, and 15 MHz; 2.5 and 20 MHz are available in one or two locations. Because there are several stations transmitting simultaneously on each frequency, you need to know the criteria for identification (modulation and schedule) in order to ascertain the usable propagation information from each location. Although these frequencies aren't near most Amateur frequencies, they are near enough to be translatable to the ham bands by interpolating between the several widely spaced frequencies available.

Canada and Russia have transmissions useful for the evaluation of propagation on several frequencies which span some of the ham bands, but they aren't the standard frequencies of worldwide coordination. For example, Canada's CHU (Ottawa) frequencies are 3330, 7335, and 14670 kHz and Australia's VNG frequencies are 4500, 7500, and 12000 kHz. These frequencies are sufficiently close to the 80, 40, and 20-meter ham bands. Russia's frequencies for RID (Tashkent) and RWM (Moskva) are 4 kHz higher and 4 kHz lower, respectively, than the standard frequencies. These systems give the DXer some propagation beacon options to enhance his operation. Next month I'll discuss some low cost and high tech monitoring equipment and its use.

Last-minute forecast

The best time for long skip openings on the higher frequency bands (10 to 30 meters) is the first few days of the first week and the fourth and last weeks of the month. Solar flux is expected to be high and, consequently, raise the MUF at this time. If the flux peaks very high (greater than 250 units), signal levels may be below normal on the lower frequencies. Transequatorial openings should be good some evenings, particularly on days of geomagnetic disturbance. Geomagnetically disturbed days are expected around the 1st, 10th, 20th, and 28th. The latter two dates are disturbances most likely associated with the starting of the winter absorption anomaly (STRATWARM). The lower bands are also affected, but should be best the second week. Expect low thunderstorm noise and good signal levels then.

Lunar perigee occurs on the 8th; a full moon appears on the 11th. The Quadrantids — a short, but intense, meteor shower — will occur between January 2 and 4, and last a few hours.

Band-by-band summary

Ten and 12 meters, the highest daytime DX bands, are nearest the MUF for paths to the Southern Hemisphere. They will be open most days during the five hours before and seven hours after local noon for solar flux levels above 220, with shorter hours for lower flux. These bands open on paths toward the east in the morning and close toward the west in the evening. The paths are up to 3500 km (2100 miles) in a single hop and multiple hops are usually available. On occasion, transequatorial openings (multiple hops without ground reflection) produce high strength signals during late evenings.

Fifteen and 17 meters are open most days to the south. They are the transition bands available to the east, or the highest to the west and occasionally to the north when the solar flux is up above 220. When used to the south, these bands have more signal distortion (multipath) than 10 and 12. The exception occurs when the bands are just opening or closing into the night, particularly on 17 meters. They are best used in other directions when the MUF is just above them.

Twenty and 30 meters are now both daytime and nighttime DX bands. Twenty is the lowest band available to

JANUARY

WESTERN USA

GMT	PST	N Asia Far East	NE Europe	E S. Africa	SE S. America	S Antarctica	SW New Zealand	W Oceania Australia	NW Japan
0000	4:00	30	30	12	10	10	10	10	12
0100	5:00	17	30	15	10	10	10	10	12
0200	6:00	17	30	15	15	10	10	10	15
0300	7:00	20	30	17	15	10	10	10	17
0400	8:00	30	30	17	15	12	10	15	20
0500	9:00	30	30	17	17	12	10	15	20
0600	10:00	30	30	17	17	15	10	15	30
0700	11:00	30	30	17	17	15	10	17	30
0800	12:00	30	30	17	17	15	12	17	30
0900	1:00	30	30	17	17	17	12	17	30
1000	2:00	30	30	20	17	17	15	17	30
1100	3:00	30	30	20	20	17	15	17	30
1200	4:00	30	30	20	20	20	15	17	30
1300	5:00	30	30	12	12	17	17	20	30
1400	6:00	30	17	10	10	17	17	20	30
1500	7:00	30	15	10	10	12	17	20	30
1600	8:00	30	15	10	10	10	15	15	30
1700	9:00	30	15	10	10	10	12	15	30
1800	10:00	30	15	10	10	10	12	15	30
1900	11:00	30	15	10	10	10	10	12	30
2000	12:00	30	17	10	10	10	10	12	17
2100	1:00	30	30	10	10	10	10	10	15
2200	2:00	30	30	10	10	10	10	10	12
2300	3:00	30	30	12	10	10	10	10	12

MID USA

GMT	MST	N Asia Far East	NE Europe	E S. Africa	SE S. America	S Antarctica	SW New Zealand	W Oceania Australia	NW Japan	CST
0000	5:00	30	30	15	12	10	10	10	15	6:00
0100	6:00	30	30	15	15	10	10	10	17	7:00
0200	7:00	17	30	17	15	10	10	10	20	8:00
0300	8:00	17	30	17	15	12	10	15	20	9:00
0400	9:00	20	30	17	15	12	10	15	30	10:00
0500	10:00	30	30	17	17	15	12	15	30	11:00
0600	11:00	30	30	17	17	15	12	17	30	12:00
0700	12:00	30	30	17	17	15	15	17	30	1:00
0800	1:00	30	30	17	17	17	15	17	30	2:00
0900	2:00	30	30	20	17	17	17	17	30	3:00
1000	3:00	30	30	20	17	17	17	17	30	4:00
1100	4:00	30	30	20	20	17	17	20	30	5:00
1200	5:00	30	20	12	20	17	17	20	30	6:00
1300	6:00	30	17	10	12	17	20	20	30	7:00
1400	7:00	20	15	10	10	12	15	15	30	8:00
1500	8:00	20	15	10	10	12	15	15	30	9:00
1600	9:00	30	15	10	10	10	12	15	30	10:00
1700	10:00	30	15	10	10	10	12	15	30	11:00
1800	11:00	30	15	10	10	10	10	12	30	12:00
1900	12:00	30	17	10	10	10	10	12	30	1:00
2000	1:00	30	17	10	10	10	10	10	30	2:00
2100	2:00	30	30	10	10	10	10	10	15	3:00
2200	3:00	30	30	10	10	10	10	10	12	4:00
2300	4:00	30	30	12	10	10	10	10	12	5:00

EASTERN USA

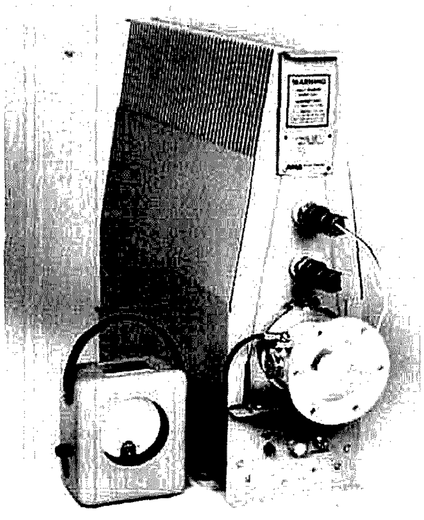
GMT	EST	N Asia Far East	NE Europe	E S. Africa	SE Caribbean S. America	S Antarctica	SW New Zealand	W Oceania Australia	NW Japan
0000	7:00	30	30	15	12	10	10	10	20
0100	8:00	30	30	17	12	12	15	15	20
0200	9:00	30	30	17	15	12	15	15	30
0300	10:00	30	30	17	15	15	15	15	30
0400	11:00	30	30	17	15	15	17	17	30
0500	12:00	30	30	17	17	15	17	17	30
0600	1:00	30	30	17	17	17	17	17	30
0700	2:00	30	30	17	17	17	17	17	30
0800	3:00	30	30	20	17	17	17	17	30
0900	4:00	30	30	20	17	17	17	17	30
1000	5:00	30	20	12	17	17	20	20	30
1100	6:00	30	17	10	12	17	20	20	30
1200	7:00	17	15	10	10	17	20	20	30
1300	8:00	20	15	10	10	12	15	12	30
1400	9:00	20	15	10	10	12	15	12	30
1500	10:00	30	15	10	10	10	15	15	30
1600	11:00	30	15	10	10	10	15	15	30
1700	12:00	30	15	10	10	10	12	20	30
1800	1:00	30	17	10	10	10	10	15	30
1900	2:00	30	17	10	10	10	10	12	30
2000	3:00	30	30	10	10	10	10	10	30
2100	4:00	30	30	10	10	10	10	10	17
2200	5:00	30	30	15	10	10	10	10	15
2300	6:00	30	30	15	10	10	10	10	17

the south at night and generally to the north in the daytime. Thirty is the main band to the north both day and night, and is the highest band open much of the time at night to the northeast and northwest. These paths may be affected occasionally by 10 to 20 dB of extra absorption from the winter absorption anomaly.

Forty, 80, and 160 meters (the night only DX bands) exhibit short skip propagation during daylight hours, then lengthen skip at dusk. These bands follow the darkness path, opening to the east just before local sunset, swinging more to the north-south near midnight, and ending up in the Pacific areas for a few hours before dawn. These bands can be used to the south, but will exhibit increased signal distortion from multipath. The winter absorption anomaly also affects these bands greatly.

High Power Absorption Wattmeters From Bird

Bird Electronic Corporation has a complete line of high power TERMALINE® absorption wattmeters with power ratings from 1.5 to 10 kW and calibrated frequency ranges of 54 to 890 MHz.



The new Bird high power models feature full capacity loads with built-in wattmeters that use temperature-compensated elements for accuracy. The units are available with 3 1/8" EIA flanged and unflanged connectors requiring no extra wattmeter line sections. All power levels (except the 10-kW models) are available with both convection and blower forced air cooling; the 10-kW versions have forced air cooling only.

All models include a temperature sensitive thermoswitch for air overheating interlock protection. Forced air units include a thermoswitch that turns on the blower automatically when needed. A manual control is provided as well. All models have an accuracy of ±8 percent full scale.

For complete details contact Bird Electronic Corporation, 30303 Aurora Road, Solon, Ohio 44139. Phone: (216)248-1200.

Circle #311 on Reader Service Card.

Custom Call Signs

SIGN ON of Merrick, New York adds a new look to their product line of transferable Custom Call Signs. The flexible plastic 2-1/4" × 8" signs now feature the words "Amateur Radio" and your call in white lettering. They may be ordered in your choice of black, blue, or red background colors.

SIGN ON products are designed for all types of automotive vehicles. A suction cup version allows for inside window mounting; the magnetic mount design attaches to exterior metal vehicle panels.

The price is $8.50 per sign, postpaid, or two for $15. Volume discounts are available. For more information contact SIGN ON, 1923H Edward Lane, Merrick, New York 11566.

Circle #312 on Reader Service Card.

New Hamtronics® Catalog

Hamtronics, Inc. has a new 40-page catalog. New products include a computerized repeater, 900-MHz transmitter, and subaudible tone decoder (CTCSS) module. A large selection of VHF/UHF transmitter autopatches, converters, preamps, and accessories are also featured.

For your copy write to Hamtronics, Inc., 65-F Moul Road, Hilton, New York 14468-9535. Phone: (716)392-9430. FAX: (716)392-9420.

Circle #310 on Reader Service Card.

New Dual Bander from Kenwood

The new Kenwood TM-701A 2 meter/70cm FM dual bander has 20 memory channels, 25 watts on 2 meters and 70 cm, selectable full duplex operation, bright amber LCD display, and more.

Features include:

- Extended frequency coverage on 2 meters
- New DTMF/control function microphone
- Dual digital VFOs
- Adjustable frequency step selection
- CTCSS tone encoder built in (optional TSU-6 enables tone decode)
- Multimode scanning with carrier or time-operated stop.

See your authorized Kenwood Amateur Radio dealer for details or write: Kenwood USA Corporation, PO Box 22745, 2201 E. Dominguez Street, Long Beach, California 90801-5745. Suggested retail price: $599.95.

AEA PakMail Mailbox/Upgrade, New PK-232MBX

Advanced Electronics Applications announces an upgrade for the PK-232 multimode data controller which includes PakMail mailbox with third-party traffic. This upgrade, the PK-232MBX (MailBox), includes firmware and a daughter board. Customers who purchased the PK-232 on or after September 15, 1989 can obtain the update *free* for a $5 shipping and handling fee.

Available to current PK-232 owners, the firmware upgrade with PakMail and daughter board is $65; the firmware only, without PakMail and TDM, is $30.

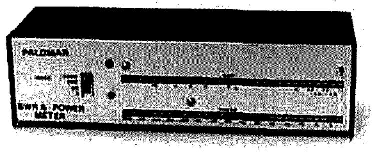
The price of the PK-232MBX will remain at the current Amateur net of $349.95. To keep the price at this level, AEA will include an RS-232 interface cable with each unit instead of the more expensive "Y" cable. The "Y" cable can be purchased as an option for a suggested retail price of $40.

For details contact Advanced Electronic Applications, Inc., PO Box C-2160, Lynnwood, Washington 98036. To order the upgrade call AEA at (206)775-7373.

Circle #308 on Reader Service Card.

SWR And Power Meter

Palomar Engineers has announced a new deluxe SWR and power meter, Model M-835 uses Palomar's patented system that gives readings continuously even when you're operating CW or SSB. It has four power ranges: 2, 20, 200 and 2000 watts. Both power and SWR



(continued on page 90)

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are displayed on 6" light bars with 3 percent resolution. It operates from 1.8 to 30 MHz and requires 12 volts DC power.

The meter sells for $189.95. An AC adapter is $15. A free catalog is available from Palomar Engineers, PO Box 455, Escondido, California 92025. Phone: (619)747-3343.

Circle #309 on Reader Service Card.

ICOM Announces New IC-2SAT, IC-3SAT, and IC-4SAT

ICOM introduces its new IC-2SAT 2-meter mini-handheld transceiver, IC-3SAT 220-MHz FM transceiver, and IC-4SAT 440-MHz mini-handheld transceiver.

The IC-2SAT has an easy access keyboard and built-in rechargeable batteries. Features include:

- 5 watt power output at 13.8 volts DC
- Built-in NiCd batteries
- 48 memory channels
- Quick tuning control
- DTMF code memory
- Built-in clock
- External DC power jack

The IC-2SAT also has auto power off timer function, power saver function, and set modes for numerous settings. Other features include: ICOM scan functions, priority watch, memory masking, memory transfer, and tone squelch. Battery packs and a battery case, chargers, headsets, and other options are also available. The suggested retail price is $439.

The IC-3SAT 220-MHz FM transceiver features:

- Built-in NiCd batteries
- Tuning control and keyboard
- 48 memory channels and one call channel
- DTMF code memory for auto dialing
- Scan functions
- Built-in clock
- External DC power jack with charging capability
- 5 watt output power at 13.7 volts DC.

IC-3SAT options include the UT-49 DTMF decoder unit, UT-50 tone squelch unit, and the UT-51 programmable tone encoder unit. Suggested retail price is $449.

The IC-4SAT is a multifunctional handheld featuring:

- 5 watt power output at 13.8 volts DC
- Built-in clock
- Built-in batteries
- Keyboard and tuning control
- 48 memory channels
- DTMF code memory

Other features include: an external DC power jack, handy repeater functions, priority watch, and a variety of scan functions. Additional options include the UT-50 tone squelch unit, UT-49 DTMF decoder unit, and the UT51 programmable tone encoder unit. Suggested retail price is $449.

For more information contact ICOM America, Inc., 2380 116th Avenue N.E., PO Box C-90029, Bellevue, Washington 98009-9029.

Circle #306 on Reader Service Card.

Picture Perfect Video Digitizer

MFJ Enterprises, Inc. announces the new MFJ-1292 "Picture Perfect" Video Digitizer for IBM compatible computers.

The MFJ-1292 is an expansion card for IBM or compatible computers that lets you plug a camcorder or video camera into your computer and use it to capture digitized video snapshots on floppy or hard disks. Display these digitized video snapshots on your computer screen and use your drawing or paint program to add lettering, color, graphics, or other enhancements.

MFJ-1292 "Picture Perfect" comes with a complete software package plus utilities. One program lets you capture digitized snapshots in VGA, EGA, CGA, Hercules or Raw Data formats. You also get programs that let you view your pictures, capture graphics from your screen into the MFJ-1292 format, convert graphics files to MFJ-1292 format, and transmit your digitized snapshot files using your MFJ-1278, MFJ-1270B, MFJ-1274, or other packet radio controller. A contrast and brightness control unit is included for fine tuning.

The MFJ-1292 sells for $199.99. For more information contact MFJ Enterprises, Inc., PO Box 494, Mississippi State, Mississippi 39762. Phone: (601)323-5869, FAX: (601)323-6551, or order toll free by calling (800)647-1800.

Circle #305 on Reader Service Card.

Cushcraft R5 No-ground Radial Vertical

Cushcraft introduces its new R5 No-ground Radial Vertical for 10, 12, 15, 17, and 20 meters. The new R5 is the third generation of Cushcraft's half-wavelength no-ground radial vertical antenna.

The R5 has optimum current distribution for low angle radiation and DXing. The antenna is only 17 feet high. It can be used for portable or fixed operation and weighs just 9 pounds.

Automatic frequency selection of all five bands is accomplished with high Q traps and a broadband solid-state impedance matching network that accepts 50-ohm input through a PL259 connector.

The R5 is available through Amateur dealers worldwide. For details contact Cushcraft Corporation, PO Box 4680, 48 Perimeter Road, Manchester, New Hampshire 03108. Phone: (603)627-7877.

Circle #304 on Reader Service Card.

Update to Russian Phrases for Amateur Radio

W6HJK has compiled a new 20-page syllabus for *Russian Phrases for Amateur Radio.* A 90-minute audio cassette has also been added to help with pronunciation.

The booklet provides English words and phrases for QSOs, accompanied by the Russian translation and the English transliteration. Sections on the Russian alphabet, phonetics, CW characters, numerals, and given names are included, along with suggestions for addressing mail to the Soviet Union.

For more information write to: Russian Phrases For Amateur Radio, Len Traubman, W6HJK, 1448 Cedarwood Drive, San Mateo, California 94403.

Circle #303 on Reader Service Card.

ICOM Announces New IC-726 HF/50-MHz All Mode Transceiver

ICOM introduces the new IC-726. This small, lightweight HF transceiver allows band operation from 500 kHz to 30 MHz and 50 to 54 MHz. Features include:

- 100 watts power output
- Wide dynamic range
- 26 memory channels
- All mode operation
- Band stacking registers

The IC-726 also includes 10-Hz digit readout, a variety of scan functions, and CI-V system for computer control. Options include an HF automatic antenna tuner and CW narrow filter. The suggested retail price is $1299.

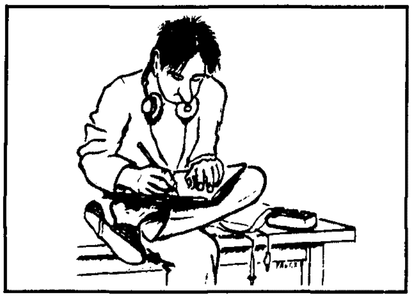
For more information contact ICOM America, Inc., 2380 116th Avenue N.E., PO Box C-90029, Bellevue, Washington 98009-9029.

Circle #307 on Reader Service Card.

Elmer's Notebook

By Tom McMullen, W1SL

ELEMENTARY ELECTRONICS — EXPLORING RESISTANCE



Resistance in an electrical circuit decreases the flow of electrons. To be more specific, resistance regulates electron flow. Resistance is one of the most useful tools available for controlling how many electrons flow and where they go. Also, voltage can be developed across a resistance, and a series of resistances can serve as a voltage divider. The list of resistance applications is long; I'll examine a few of them.

Electron movement

In order to appreciate what's happening in a resistor, you need to take a look at some basic electron theory. You've been told that: (a) electrons orbit around a nucleus in an atom, (b) electrons are very difficult to dislodge from an orbit, and (c) when dislodged, electrons move at nearly the speed of light. Experience with a multitude of electronic devices tends to confirm this theory.

Some materials — like lead, copper, silver, and gold — conduct electricity very well. This indicates that their electrons are easily dislodged from orbit. Other substances — like carbon, steel, nickel, and chromium — conduct poorly. Still other materials — like glass, porcelain, and most plastics — don't allow any current flow to speak of. This means it's a pretty good general assumption that the softer metals (copper, silver) have cooperative electrons, the harder metals (steel, chromium, carbon) have electrons that tend to stay in orbit, and the really hard materials (glass, porcelain) have electrons that are almost impossible to dislodge. Plastics, although not physically hard, have a specialized molecular structure that makes them good insulators.

Moving the electrons

When you connect a source of power (like a battery) to a conductive circuit (a wire), the excess electrons available at the negative (−) terminal try to get to the other, or positive (+), terminal — which has a scarcity of electrons. Imagine that you can see the first atom in a piece of wire, with all its electrons spinning about the nucleus. Along comes an electron from the battery, pushing its way into the crowd. There's only room for a specific number of electrons in each atom (described in the atomic tables found in physics reference books), so in order to let this new electron in, one must be bumped out of place. Because the electrons in the outer orbit have less physical attraction to the nucleus than those in the inner orbits, one of the electrons in the outer orbit is dislodged. It's interesting to note that the good conductors (copper, silver, gold) all have just one electron in their outermost orbit (or shell). This electron is called a "free" electron because it's more easily dislodged from orbit than the electrons in the inner shells. The electrons of the inner shells are called "bound" electrons and are extremely difficult to dislodge.

Because an electron without a home won't survive very long, the bumped electron immediately dashes over to the atom next door and joins it. This dislodges an electron from that atom, which moves to another atom, and so on, down the length of the wire (see **Figure 1**). Eventually the last atom in the wire gets bumped, and the most recently dislodged electron finds a home in the (+) terminal of the battery.

This process continues until the imbalance of electrons in the battery is neutralized and no more current (electrons) will flow. At this point, the battery is "dead." (Other power sources, like rectifiers, also provide excess electrons and accept electrons from the circuit, but it's easier to work with a simple cell or battery when exploring basic theory.)

Even though this representation deals with one atom at a time, remember that there are millions of atoms in a piece of wire, and that this action is taking place in all of them at the same time — causing a great number of electrons to move at once. In fact, the description of a Coulomb (C) involves the number of electrons that pass a given point in one second:

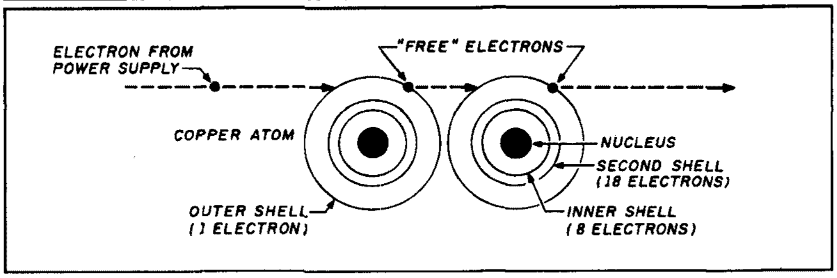
1C = 6.28×10^{18} electrons per second

This also is a current of one ampere.

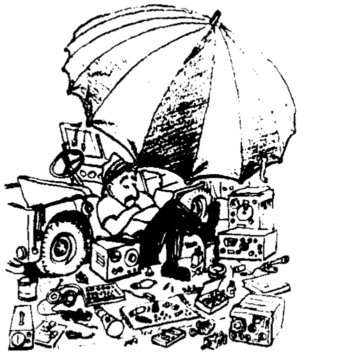
Controlling the flow

To put the electron flow to use, you must be able to control the amount of current flowing in a circuit. If all the electrons available at the negative terminal of the supply were to rush

FIGURE 1



In this simplified representation of copper atoms, an electron from an external power source joins the outer orbit (shell) of an atom, dislodging its "free" electron. That electron migrates to the next atom, and so on. This action is repeated for the length of the wire as long as the power source povides extra electrons.



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DEADLINE 15th of second preceding month.

SEND MATERIAL TO: Flea Market, Ham Radio, Greenville, N. H. 03048.

BEGINNER'S RADIO CLEARINGHOUSE. On a space available basis, we are going to offer you, OUR SUBSCRIBER, free of charge, a chance to find a home for your used equipment with a new Ham. Please send us a short description of what you want to sell along with price, name, address and phone number. We'll run it once in a special section of the classified ads under the heading of BEGINNER'S RADIO CLEARINGHOUSE. Please limit your ad to 20 words or less.

SWAN 500 CX/117XC TRANSCEIVER/PS $250. Kenwood TS-820S Transceiver $450. David Gater, WD4MWR, 1000 Tarpon Ct, Nokomis, FL 34275. (813) 485-8937.

10 Mtr FM Hy-Gain CB Board with instructions $9.95. FM DET Kit $7.00. 40 channel switch $5.00. Crystal $5.00. SQ/VOL pots $2.00/pair. All for $24.95 plus $2.00 s/h. MORNING DISTRIBUTORS, PO Box 717, Hialeah, FL 33011. (305) 884-8686.

FOR SALE: DX100, NC300. Both excellent condition and operational. W3SAA. (215) 332-8998.

VHF/UHF MODULES for 6M to 23cm. Catalog. TEC, PO Box 1743, Melbourne, FL 32902. (407) 676-6907.

WANTED: Pre-1942 TRANSMITTER. Any condition or part considered. Also Hammarlund SuperPro SP-10 thru SP-220. Bob Mattson, KC2LK, 10 Janewood, Highland, NY 12528. (914) 691-6247.

STOP BEING STARTLED by those high volume transmissions! The REGEN-I module automatically maintains the voluma level oa amateur, business (police, fire, power companies, fleet operators, etc.), CB, marine radios and scanners. The **REGEN-I** will amplify low signals as well as attenuate a strong audio signal to keep volume at a nearly constant level regardless of incoming signal strength. The REGEN-I amplifies the low level signals up to 100 times and reduces un-squelched FM hiss. The REGEN-I eliminates the need to constantly manipulate the volume control to compensate for changing signal strength and modulation levels. Quietly, instantaneously, with no turn on delay, amplifies and attenuates! Ideal choice for Base and Mobile Transceivers and Scanners where a constant receiver audio level is important for safety and convenience. Compatible with sets using AUDIO ATTENUATOR type volume controls (Most transceivers in use!). Operates from the radio's power source. 8 to 35 VDC ª 10 ma. Simple instruction included for installation; attaches with adhesive foam backing. PRICE: $49.95, includes shipping. $10.00 off for prepaid orders received in the mail or by FAX (Credit card orders)! Must refer to this ad! Master Card/VISA, MO, Checks OK. No cash or COD's please. EMULATION ASSOCIATES, 520 Glenbrook Road, Suite 203-A, Stamford, CT 06906. *SALES: 203-356-1632. FAX: 203-323-9044.*

HAM SOFTWARE IBM/Compatibles 10 disks $26.95. MC/VISA/Discover. N5ABV EAPCO/H, Keller, TX 76248-0014. (817) 498-4242. 1-800-869-7208.

CHASSIS AND CABINET KITS. SASE. K3IWK, 5220 Harmony Grove Rd, Dover, PA 17315.

PK232 QUICK REFERENCE. All commands through the October 1989 firmware revision functionally grouped. Send $4.95 to Bryan R. Leipper, K1CD, 714 Terra Court, Reno, NV 89506.

POLICE/FIREFIGHTER HAMS. Please send your call, name, address, rank, department name for inclusion in special roster, available early 1990. Capt. Bob Blakeslee, N2IHQ, 1-1/2 Macomber Avenue, Binghamton, NY 13901.

WANTED: Technical information on Rubidium Frequency Standards made by Collins/GenRad (O-1622/AFS-81). Also interested in spares and parts for two units. Kirk Bailey, N7CCB, PO Box 1702, Corvallis, OR 97339. (503) 753-9051.

RTTY JOURNAL published 10 times per year for those interested in digital communications. Read about RTTY, AMTOR, MSO'S, PACKET, RTTY DX and Contesting. Plus technical articles concerning tne digital modes. $12.50 per year (foreign higher). RTTY JOURNAL, 9085 La Casita Ave, Fountain Valley, CA 92708.

CUSTOM MADE EMBROIDERED PATCHES. Any size, shape, colors. Five patch minimum. Free sample, prices and ordering information. HEIN SPECIALTIES, Inc., 7960 SW Manitou Trail, Glen Arbor, MI 49636. (616) 334-4385.

IBM-PC RTTY/CW. New CompRtty II is the complete RTTY/CW program for IBM-PC's and compatibles. Now with larger buffers, better support for packet units, pictures, much more. Virtually any speed ASCII, BAUDOT, CW. *Text entry* via built-in screen editor! Adjustable split screen display. Instand mode/speed change. Hardcopy, diskcopy, break-in buffer, select calling, text file transfer, customizable full screen logging, 24 programmable 1000 character messages. Ideal for MARS and traffic handling. Requires 256k PC or AT compatible, serial port, RS-232C TU. $65. Send call letters (including MARS) with order. David A. Rice, KC2HO, 144 N. Putt Corners Rd, New Paltz, NY 12561.

LET THE GOVERNMENT FINANCE your small business. Grants/loans to $500,000. Free recorded message: 707-448-0270. (KH5).

SLEP SPECIALS: *Plate transformers for amplifier builders or* replacement, Gonset P/N 271-107 for Models 903, 913 linear amplifiers, 115VAC/1850V at 500MA, size 5-1/4L x 4-1/4W x 4-1/2H, Wt 18 lbs, new $37.00, Military USM-207 frequency counter, late model solid state, eight digit readout, 0-500 MHz, high stability crystal oven oscillator, size 19"W x 5"H x 17"D, Lab quality tested $185.00, Military URM-25H portable *signal generator, 10 kHz thru 50 MHz, calibrated output, ideal* for radio repair $145.00, have quantity, satisfaction guaranteed, VISA/MC or check, add shipping, write/phone Bill Slep, 704-524-7519, SLEP ELECTRONICS COMPANY, Highway 441, Otto, NC 28763.

BEAT THE COLD! Melbourne, FL QTH: 4:2:2 (3100 sq/ft) house, 1.6+ acres, 220V wired shack with coax races built in, RHON/25 pad, workshop, pool, sprinkler system, in the country no restrictions, many tall pines and oaks, easy commute to Cape Canaveral, close to fishing/beaches/shopping, SAE for details and photos, $209,000. NODH/4.

DIGITAL AUTOMATIC DISPLAYS. Any Radio. Be specific. Large 45 cent SASE. GRAND SYSTEMS, POB 2171, Blaine, WA 98230.

AVANTEK ATF10135, $12.00, MMIC's, P.C. board, SASE: WA3IAC, 7148 Montague St, Philadelphia, PA 19135.

"HAMLOG" COMPUTER PROGRAM. Full features, 17 modules. Auto-logs, 7-band WAS/DXCC. Apple $19.95. IBM, CP/M, KAYPRO, Tandy, C128 $24.95. HR-KA1AWH, POB 2015, Peabody, MA 01960.

TEST EQUIPMENT—reconditioned (guaranteed), terrific prices, NBS *Calibration* available. VISA, M/C, Amex, COD. ELECTRONIC SURPLUS, INC, 4350 Town Plaza, Suite 203, Houston, TX 77045. (713) 728-9718.

WANTED: Dentron CM-U. N0BUC, 118 Elizabeth Drive, Aberdeen, SD 57401.

YAGI BUILDERS. 6061-T6 tube traps. Good for 1500 PEP. SASE for details. No collect calls. BROWN ENGINEERING INC, 5501 SW 25th Court, Hollywood, FL 33023. (305) 989-4658.

FREE Ham Gospel Tracts, SASE, Steve Forst, N3FTT, 5133 Gramercy, Clifton Heights, PA 19018.

WANTED: Ham equipment and other property. The Radio Club of Junior High School 22 NYC, Inc. is a nonprofit organization, granted 501(C)(3) status by the IRS, incorporated with the goal of using the theme of Ham Radio to further and enhance the education of young people nationwide. Your property donation or financial support would be greatly appreciated and acknowledged with a receipt for your tax deductible contribution. If you call or write today it's not too late to get a 1989 tax deduction. It's easier, faster, and usually more profitable to donate than to sell. Most important, you're helping. Write us at: PO Box 1052, New York, NY 10002. Round the clock hotline: (516) 674-4072.

450 MHz SPECTRUM ANALYZER. Adapted from Nov 85 QST article by Al Helfrick, K2BLA. Features: Three digit LED center frequency digital readout. 12 position calibrated Scan Width, 1 kHz to 50 MHz, switchable bandwidth; Wide ⁵ 300 kHz, Narrow ⁵ 10 kHz. Use your low frequency scope for the display portion. Switchable 10 kHz video filter. Variable IF attenuator, LOG output calibrated in 10 db steps. For complete kit, order #450-KIT $459.95 plus $4.50 s/h. For assembled and tested unit, order 450-AST, $799.95 plus $6.50 s/h. Calif residents add 6% sales tax. Foreign orders add 15% for shipping. A & A Engineering, 2521 W. LaPalma, #K, Anaheim, CA 92801. (714) 952-2114.

IMRA International Mission Radio Association helps missionaries. Equipment loaned. Weekday net, 14.280 MHz, 1-3 PM Eastern. Nine hundred Amateurs in 40 countries. Rev. Thomas Sable, S.J., University of Scranton, Scranton, PA 18510.

FOR SALE: Kenwood R-2000 SW receiver, $400. Ranger AR-3500 10 meter, 100W mobile XCVR, $300. Both excellent condition and appearance. Matt, WA1HRE, (203) 693-0468/693-6596. Leave message.

1989 CALLBOOK - US $14, foreign $17. Only have one copy each, include SASE in case already sold. Nate Williams, W9GXR, 6915 Prairie Drive, Middleton, WI 53562.

INTERESTED IN PUBLIC SERVICE? Join your Local Radio Emergency Associated Communications Team. In Pennsylvania call (717) 938-6943 or write REACT, 1160 Old Trail Rd, Etters, PA 17319.

ANALOG AND RF CONSULTING for the San Francisco Bay area. Commercial and military circuits and systems. James Long, Ph.D., N6YB (408) 733-8329.

RUBBER STAMPS: 3 lines $5.00 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

ELECTRON TUBES: Receiving, transmitting, microwave... all types available. Large stock. Next day delivery, most cases. DAILY ELECTRONICS, PO Box 5029, Compton, CA 90224. (213) 774-1255.

RECONDITIONED TEST EQUIPMENT $1.25 for catalog. Walter, 2697 Nickel, San Pablo, CA 94806.

COMING EVENTS

Activities — "Places to go . . ."

SPECIAL REQUEST TO ALL AMATEUR RADIO PUBLICITY COORDINATORS: PLEASE INDICATE IN YOUR ANNOUNCEMENTS WHETHER OR NOT YOUR HAMFEST LOCATION, CLASSES, EXAMS, MEETINGS, FLEA MARKETS, ETC, ARE WHEELCHAIR ACCESSIBLE. THIS INFORMATION WOULD BE GREATLY APPRECIATED BY OUR BROTHER/SISTER HAMS WITH LIMITED PHYSICAL ABILITY.

INDIANA: December 31. South Bend. Hamfest Swap & Shop, December 31, last day of 1989 BEFORE New Year's Day at CENTURY CENTER downtown on U.S. 33 ONEWAY North between Trustcorp Bank Building and river. Four lane highways to door from all directions. Tables: $5/5 ft. Round; $15/8x2.5 rectangular; $20/8 ft. Wall locations. Over half acre on carpeted floor. Electric NO charge— Bring long cord. Talk in 52, 39, 09, 94, 145.29. Call 219-233-5307, write Wayne Werts, K9IXU, 1889 Riverside Drive, South Bend, IN 46616.

FLORIDA: January 20. The 10th annual Citrus County Hamfest sponsored by the Sky High ARC, National Guard Armory, Crystal River. Open to public 9 AM. Exhibitors 7 AM. Tickets $4/advance by Dec. 20 and $5 at door. XYL's free with OM. Contact Del Slocum (904) 726-0725 or write SHARC Hamfest, 3101 E. Oakton, Hernando, FL 32642 for tickets, tables, flea market space or info.

NEW YORK: January 21. Electronics Fair and Giant Flea Market sponsored by the Metro 70cm Network, Lincoln High School, Kneeland Avenue off Yonkers Avenue, Yonkers. Setup 8 AM. Public 9 AM. Rain or shine. Admission $3. Tables $12. Contact Otto Supliski, WB2SLQ, 53 Hayward St, Yonkers, NY 10704. (914) 969-1053.

MISSOURI: January 27. The St. Louis Repeater Club's "Winterfest", Stratford House, I-44 just west of I-270. 9 AM to 3 PM. Sellers admitted 7 AM. Features VE exams, indoor flea market and refreshments. Admission $1.00 per person. $5/selling space. For reservations and information call WA0FQK. (314) 351-7732.

YOUTH LINK NET. Open to all Hams under age 18. Saturdays at 2000 UTC, 28.425 MHz. For more information contact Net Control, George Manning, WB5NMH, 602 Glendale St, Burkburnett, TX 76354.

FREE 1989-90 Florida two meter repeater directories are currently being distributed by the Hernando County Amateur Radio Assn. of Brooksville, FL. Ask for one at any official Florida Welcome Center or SASE to Repeater Directory, Hernando County ARA, POB 1721, Brooksville, FL 34605-1721.

LAUREL ARC monthly (except December) Amateur exam sessions for all license classes. No fee is charged. Pre-registration is required. Call (301) 725-1212, Maryland Radio Center, 8576 Laureldale Drive, Laurel, MD 20707.

AMATEUR RADIO CLASSES: For those people interested in obtaining a Novice (basic level) Ham license or upgrading to Tech/General, the Chelsea Civil Defense, in cooperation with QRA Radio Club, will sponsor Amateur Radio Communications classes evenings at Chelsea High School starting MARCH 7, 1989. For more information write Frank Masucci, K1BPN, 136 Grove Street, Chelsea, MA 02150. Please enclose your telephone number.

through the circuit immediately, you would quickly have a dead battery, and the wire would probably vaporize in a bright flash (more on this later).

Reducing the size of the wire is one method of controlling current flow, but this really isn't practical. A piece of wire the size of a hair still contains several million atoms in a cross section. By selecting your materials carefully, you can build a current-restricting device. Interestingly enough, it's called a resistor because it resists the flow of electrons.

One of the most common resistor materials is carbon — a very hard substance. By carefully controlling the mixture of carbon with other materials, and the size and shape of these materials as they are combined in the resistor, you can tailor resistance to suit a particular need. Smaller numbers of electrons moving through a circuit will be admitted into a carbon-composition structure, allowing fewer electrons to flow at any given time.

Early resistors were simple rods made of carbon and some type of cementing material. Some were 1/4 inch in diameter and 1 or 2 inches long; some were the size of the lead in a pencil and 1/2 inch or less in length. These carbon rods had wires wrapped around their ends and were glued in place with conductive paste. Later models had a similar rod structure, but used a crimped brass or copper cap on the end; wires were either welded or soldered to the caps. Modern day resistors come in a variety of sizes — some are still 1/4 inch or more in diameter, while others are tiny chips less than 0.1 inch square. There are also variations in the materials used. Carbon-composition resistors are still the most widely used type (see Figure 2). There are also metal film resistors made of a very thin metallic film deposited on a ceramic or glass base with attached leads. Some resistors are made out of ceramic tube covered completely with resistive material, with portions of the material etched away in a spiral path to obtain the desired resistance.

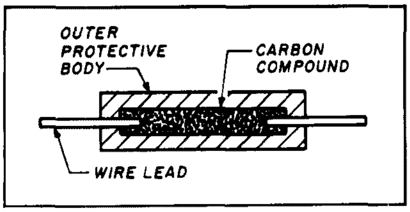
Using resistance

Resistance has so many uses that I can't begin to discuss them all here. Let's look at some of the most common ones.

Current limiting. A resistor placed in an electrical path limits the number of electrons (current) that can flow, thus protecting the device in the circuit from being overheated by excess current.

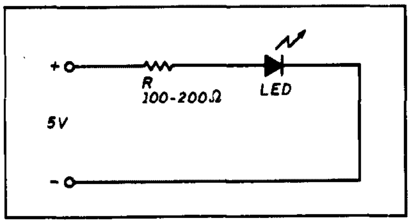
A good example of this situation is the resistor that's almost always used in series with an LED indicator (Figure 3). The LED, being a diode, will conduct at approximately 0.45 volts. But most of the circuits using LEDS have a 5-volt supply. The series resistor limits the current that can flow in the diode, and also drops the supply voltage to the level the diode needs to conduct and illuminate. Transistors and vacuum tubes require a similar protection scheme to prevent the burnout that results from too much current flow.

FIGURE 2



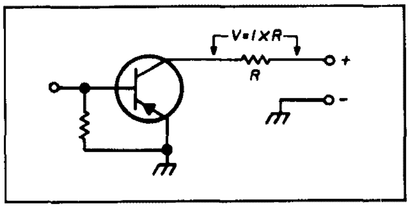
A cut-away view of a carbon-composition resistor. The composition determines the resistance and the physical size determines the heat (power) dissipation capability of the resistor.

FIGURE 3



A resistor in series with an LED serves to limit the current flow through the LED, thus preventing the diode from burning out.

FIGURE 4



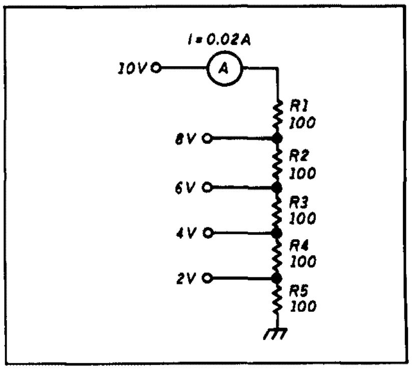
A resistor in series with the collector of a transistor limits current flow. At the same time, it provides a voltage change that varies with the drive signal applied to the transistor base.

Current-restricting resistors can be used in another way. As shown in Figure 4, the resistor between the supply voltage and the transistor collector will limit the current flow and, in doing so, will cause a voltage difference between one end and the other Because voltage is related to current flow and resistance (Voltage = Current × Resistance), you can change the voltage across the resistor by changing the current flow. This is how a "signal" is developed. Changing the driving signal to the base causes the transistor to conduct more or less, changing the current flow and, in turn, creating more or less voltage across the resistor. This voltage change is thus a larger (amplified) replica of the input signal to the transistor.

Voltage divider. Sometimes you need a voltage less than that available from the supply. Conveniently, several resistors can be placed in series to provide a selection of voltages at each junction (see Figure 5). Note that the total current flow is determined by the total resistance in the circuit (R1 + R2 + R3 + R4, and so on), while the voltage across each resistor is determined by the current flow and the resistance of that individual resistor (I × R). Also note that the sum of all the voltages must equal the supply voltage — nothing is lost or unaccounted for.

Current divider. What can you do when the available resistors can't handle the current flow? Place two or more resistors in parallel so the current flow is divided among them (Figure 6). For instance, say you have a circuit that must dissipate 2 watts of power (Watts = Current × Voltage), but you have

FIGURE 5



Several resistors in series provide a voltage divider. Note that the voltages shown are for the resistor only. Any load connected to a voltage point will use extra current, which will decrease the voltage at that point.

only 1-watt resistors. Use two resistors of equal value in parallel. Because half the current is flowing in each resistor, each is now dissipating (passing current without overheating) 1 watt. However, in order to have the same voltage across the pair, you must use resistance values that are twice that of one resistor. This is because resistors in parallel decrease the effective resistance according to the formula:

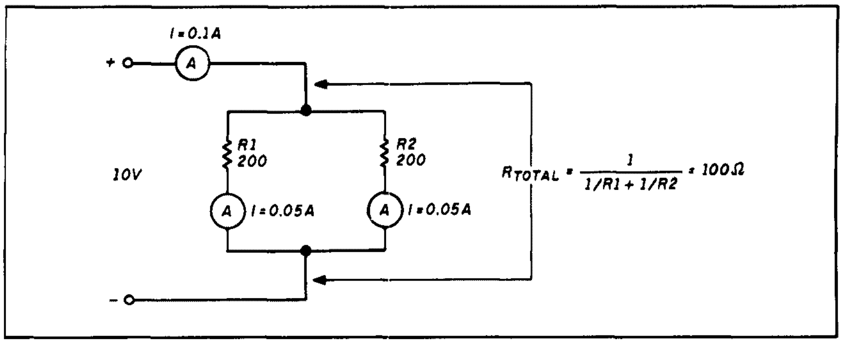
R(total) = 1/((1/R1)+(1/R2)+(1/R...)).

There are a couple of interesting rules involved here. First, the voltage is the same across all resistors. Second, if the resistors aren't the same value, the resistor with the lowest value will carry the most current (see **Figure 7**). It's also interesting to note that the total current in the circuit is the sum of all the currents in the branches: I(total) = I(R1) + I(R2) + I(R...).

The current-divider principle can provide a very accurate means of finding an unknown resistance. This is called a "bridge" circuit or sometimes a "Wheatstone" bridge (see **Figure 8**).

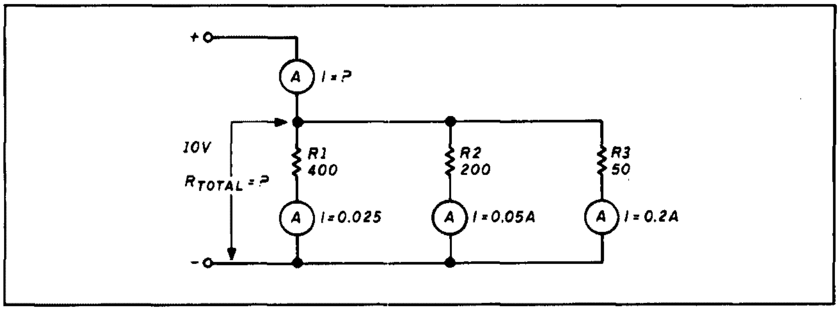
The circuit is made up of four resistors: R1 and R2, which are of the same value; R3, which is variable with a calibrated dial; and Rx, which is an unknown value. If the resistance values are the same on both sides of the circuit (R1 + R3 = R2 + Rx), then the current flow is equal for both sides. A very sensitive meter (usually a microammeter) is connected across the bridge; it won't show any reading when the two sides are balanced. If Rx is not equal to R3, then the current flow is not equal in both sides of the bridge, and the meter will deflect toward either R3 or Rx — whichever has the lowest resistance. Determine the value of Rx by changing the value of R3 until the meter shows a zero (center) reading, then reading the value of R3 from the meter's dial. This principle is used in electronic circuits from detectors to mixers to SSB generators, along with S-meter circuits, and many more. Variations of this circuit work with measuring inductance, capacitance, and impedance, and with AC or DC currents.

FIGURE 6



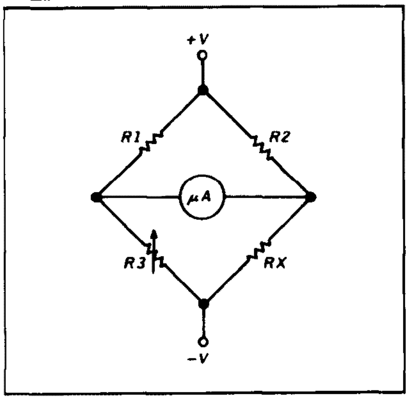
Resistors in parallel allow the current flow to be divided between them.

FIGURE 7



Unequal resistors in parallel carry unequal currents, but the total current is the sum of all the currents. Add the totals of R1, R2, and R3 to find the I(total).

FIGURE 8



A "bridge" circuit is useful for many things, including finding the precise value of an unknown resistor (Rx). See text for a description of how the circuit works.

Hidden resistance

Sometimes resistance isn't obvious. This hidden resistance may be either a help or a hindrance. Another name for hidden resistance is internal resistance. An example of "good" internal resistance is illustrated in the battery and wire experiment in last month's column. Fortunately, the battery has an internal resistance that's large enough to keep infinite current from flowing. Otherwise, the wire would have vaporized — along with the compass, your fingers, and perhaps a good chunk of the workbench. This battery internal resistance results because the chemical (electrolyte) in the battery is able to "liberate" only a limited number of free electrons and make them available to external circuitry. The down side is that this same resistance limits the available current. This means that when your circuit needs more current, you have to use a physically larger battery.

Internal resistance makes things complex in ordinary meters — like voltmeters, ammeters, and ohmmeters. These are all basically current-measuring devices, with resistance networks and calibrated scales that allow you to interpret their reading in volts, amperes, or ohms. The most common type of meter uses a very small coil of fine wire placed between the poles of a magnet. When current flows, a magnetic field is generated which reacts with the magnet. The resulting "push" between the two magnetic fields causes the coil and its attached pointer to move to the location where the force of the field is balanced with that of a small spring attached to the pointer. This small wire

has resistance which limits the current flow. This is good because it helps prevent the meter's coil from burning out. However, this resistance must be taken into account when the meter is manufactured and calibrated, and adjustments must be made in the resistor network to obtain accuracy. You, too, will have to account for this internal resistance if you ever try to recalibrate the scale of a meter by placing it in series with another well-calibrated meter. All electronic devices have internal resistance, which often provides protection against excess current flow. In most cases, this resistance is also necessary to make the device work. Just remember that it exists and allow for it in your circuitry whenever precision is vital.

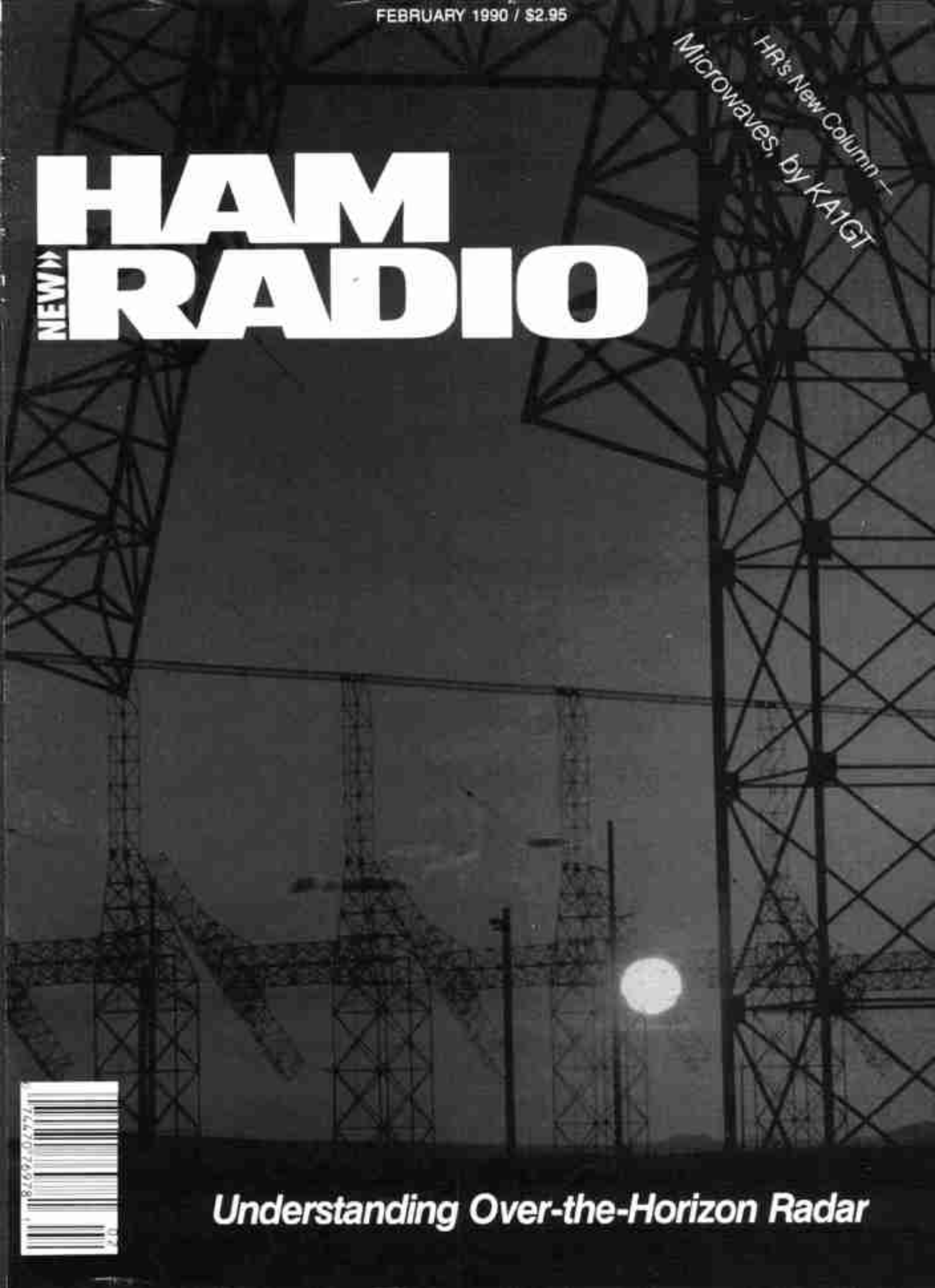
In summary, resistors play an essential part in the world of electronics and can provide material for some fascinating experiments in how electrons behave. Pick up a grab bag of resistors at your next hamfest and try some circuits — series, parallel, series and parallel, and bridge — to see what happens when you use various values and combinations. hr

FEBRUARY 1990 / $2.95

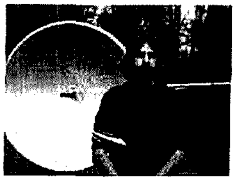
HAM RADIO

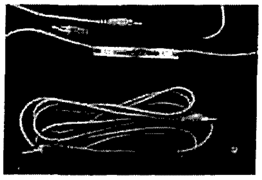
NEW

HR's New Column — Microwaves, by KA1GT



Understanding Over-the-Horizon Radar


KA1GT, page 20


NO7G, page 30

HAM RADIO

FEBRUARY 1990
Volume 23, Number 2

Cover photo: Portions of the transmit antenna arrays for the West Coast Over-the-Horizon Radar System. The antenna sections pictured are part of a nearly 4,000-foot long array located near Christmas Valley in southern Oregon. Photo courtesy of GE Aerospace, Syracuse, New York. Special thanks to the Air Force Public Affairs Office at Hanscom Air Force Base, Massachusetts for their help in obtaining the photo.

FEATURES

DEPARTMENTS

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Microfilm copies are available from Buckmaster Publishing Mineral, Virginia 23117. Cassette tapes of selected articles from **HAM RADIO** are available to the blind and physically handicapped from Recorded Periodicals, 919 Walnut Street, Philadelphia, Pennsylvania 19107.

Second-class postage paid at Greenville, New Hampshire 03048-0498 and at additional mailing offices. *Send change of address to HAM RADIO, Greenville, New Hampshire 03048-0498.*

Somewhere Over the Horizon (radar, that is)

Though it has long been one of the most insidious of the co-inhabitants of the Amateur bands, over-the-horizon (OTH) radar is also one of the most fascinating technologies in the field of electronics. There are few Amateurs, anywhere, who haven't had at least one run-in with the infamous Russian Woodpecker. But did you know that the United States, Australia, and Canada have working systems and such systems may be in the developmental stages in other countries?

Bryan Bergeron, NU1N, has written a well-documented piece on OTH radar. It appears as our feature article this month. It's interesting to note that this technology has a number of different uses. From peaceful applications like determining wind and wave conditions over the oceans to searching the skies and seas for potential military threats, OTH radars are very powerful tools.

Bergeron points out the significant potential for interference created by OTH systems. However, those in the United States have been carefully designed to minimize or eliminate this problem. And, overall, there have been few if any complaints generated by the U.S. systems in use today.

The computers that control U.S. OTH radars are preprogrammed to operate over a spectrum selected to reduce harmful interference to Amateur, broadcast, and fixed service communications. These radars can also listen in on specific frequency ranges before transmitting, to determine if there is a potential for interference. And if the frequency is in use, the system searches for another usable spot.

Many of the advances in digital signal processing have come as a direct result of OTH radar development. Computers are the single most critical ingredients of a functioning OTH radar system. The complexity of the return signal can't be processed in the same manner as a more conventional line-of-sight radar. In fact, World War II radar operators would marvel at the differences between today's OTH radar and their old PPI (cathode ray tube) displays!*

Additional advances have been made in the scientific study of our ionosphere and radio wave propagation. In order to maximize OTH efficiency, transmissions must be made near the maximun usable frequency (MUF). Taking the input from various sources, the OTH radar's transmitter frequency varies continuously depending on the best propagation path.

Finally, the United States has extensive and stringent environmental impact study regulations for radar that must be met before an operational license can be granted.

Defense has been the prime mover behind developmental work in OTH technology and hams have benefited from much of the work that has been done. I'm sure you'll find Bergeron's article fascinating reading. Perhaps you'll want to look at many of the references he's listed. Most should be available from your local library, through inter-library loans, or at the libraries of local colleges or universities.

de NX1G

* *A Race on the Edge of Time* by David E. Fisher, Ph.D. is a fascinating look at the development of radar. It's interesting to compare the OTH technology of the eighties to the "Chain Home" British radar that operated between 25 and 30 Mcs at the beginning of World War II. It's mind boggling when you contrast what was done then with what we are doing now! (If you'd like to read this book, it's available from the HAM RADIO Bookstore for $19.95 plus $3.75 shipping and handling.

Comments

AM is alive and well

Dear HR

I would like to take umbrage at the remark made by Bill Orr in the December issue, "Too bad the days of amplitude modulation are past —" They are not! If Bill would listen in his own back yard (California) he would find that AM operation is rampant. The SPAM (Society for the promotion of Amplitude *Modulation) is headed* by WB6TRQ in California. The NCS for the regular Wednesday night activity is W6RNC on 3870. Saturday evenings AM is on 7160. Almost every day that the band is open on 10 meters AM congregates around 29.000.

Besides the usual Johnson Rangers, Vikings, and Valiants, there are quite a few Collins KW-1 transmitters on the air: W6HDU, W7ISX, W7GCO, W8AHC, and WA3PUN.

QST has made similar snide remarks about AM. I hope you will not permit your writers to do the same again.

Byron H. Kretzman, W2JTP/7,
Woodinville, Washington

W2JTP/7 is the author of ***The New RTTY Handbook*** *(CQ Publishing Group, out of print), the definitive book on "green key" RTTY operation. Ed.*

First choice

Dear HR

This is my first year with *Ham Radio* magazine: what a blessing, what a change! I've seen early printings, which were fine reading. This "new" magazine is by far the best. I subscribe to others. This is the pick of the litter. Please keep up the excellent work. All your staff have shown us a super magazine, and I am sure of more great things in the future.

Congratulations to all.

V. L. Henzel, WB7QBE,
Richland, Washington

Must be a duck

Dear HR

We live in a real world. So quack, quack, quack. How quickly we forget.

OMB Director Darman stated to Congress: "It it looks like a duck, walks and talks like a duck, then it's a duck."

Take any time-worn political doublespeak like "revenue enhancement" and it still quacks like a tax duck. A tax by any other name like "user fee" is still a *tax*. Why try to convince any Congressional tax writing staff member of the extraordinary contributions of Amateur Radio over the past 70 years? When push comes to shove, the bottom line is the paranoid frenzy to get bucks — any way, any form, any pretense. We shall have user fees. So: "Read my lips, no new taxes" and be happy you are only paying user fees for licenses; of course this is *not* a tax!

As was stated a very long time ago in our history, the only time that person and property are safe is when Congress is not in session.

Gene Shapiro, W0DLQ,
Prairie Village, Kansas

Pet peeve

Dear HR

The "new" *Ham Radio* is great. Best wishes for continued success.

I'd like to air a pet peeve: Generals and above on the 28.3 to 28.5 MHz band...

I, and other Novices and Technicians, get run off the band every day by these operators. Yesterday was a perfect example: an Extra class running 500 watts, splattering over 5 kHz each way for three hours. You Generals, Advanced, and Extras can run 1500 watts and talk all over the bands; us "little guys" are limited to just 28.3 to 28.5 MHz and 200 watts PEP.

I think it's rude, inconsiderate, and not in the ham spirit for you big strappers to operate on the only 10-meter voice spectrum available to us. Yes, you *are* qualified to work there — that isn't the point. We'd love to work 29 MHz FM — that's out of band for us. How about taking your linears and privileges somewhere else?

Mike Gleeson, N6SWA,
San Diego, California

Advice needed!

Dear HR

I am about to attempt an amateur R/C airplane project, and could certainly use some expert advice from any hams out there who have experience with homebrew radio control or TNC projects. I would be happy to hear from anyone with related interests.

Dave Ewen, N0KET,
4818 New Jersey #203,
Wichita, Kansas 67210

Summer conference planned

Dear HR

The Central States VHF Society conference will be held in Wichita, Kansas during July of 1990. We are currently seeking technical programs to be presented at the conference. Programs can range from 15 minutes for a simple hint to 1 hour for a complete project. Persons wishing to present material at the conference should contact Jon Jones, N0ØY, 1116 Gatewood Court, Wichita, Kansas 67206 or John Lock, KFØM, 1427 S. St. Clair, Wichita, Kansas 67213.

John Lock, KF0M,
Wichita, Kansas

UNDERSTANDING OVER-THE-HORIZON RADAR

A glimpse at how the other half lives

By Bryan P. Bergeron, NU1N, 30 Gardner Road, Apartment 1G, Brookline, Massachusetts 02146

Have you ever wondered what the "Russian Woodpecker" is really up to? Did you know that the United States has several woodpeckers of its own, and that several more are scheduled to begin operation in the near future on frequencies that could include the Amateur bands? I'd like to examine the source of one of the most bothersome signals on the HF bands: over-the-horizon (OTH) radar.

Some radar fundamentals

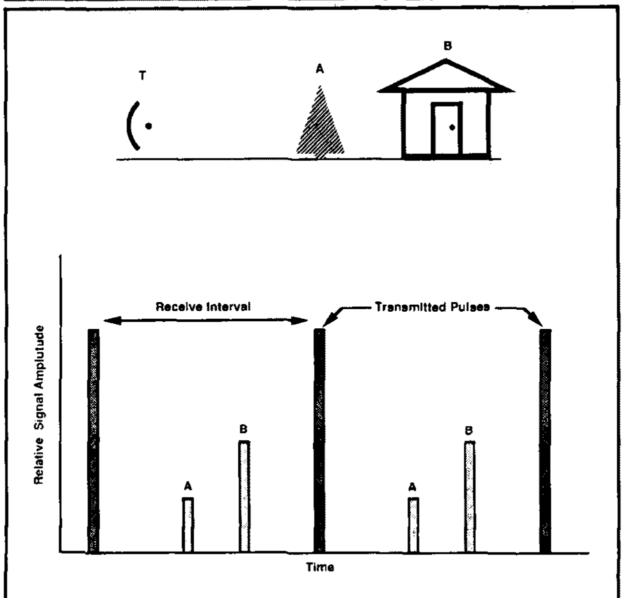
Before I get into the specifics of OTH radar, a brief review of some radar basics is in order. Radar, a term derived from the phrase **RA**dio **D**etection **A**nd **R**anging, is based on the principle that RF pulses generated by a carefully timed transmitter are reflected by the objects they encounter. The time it takes for the reflected beam to return to the transmission site can be used to determine the distance or range of the object. Radio waves travel at the speed of light, about 186,000 miles per second. For a target at a distance of 186 miles, the round trip (2 × 186, or 372 miles) would take about 2,000 μs:

$$\frac{186\ miles \times 2}{186{,}000\ miles/sec} = 2/1{,}000\ sec = 2{,}000\ \mu sec \qquad \textbf{(1)}$$

Using this relationship, you can compute relative distances accurately for any given elapsed time.

Because the frequency of the reflected signal is the same as that of the transmitted signal (assuming that there is no relative movement between the transmitter, receiver, and the reflecting object), the transmitter must be turned off for the receiver to detect the much weaker reflected signal. It follows, then, that most radar is pulsed. The transmitter sends out short intense pulses of energy with relatively long intervals between pulses, during which time the receiver is active (see **Figure 1**). When sufficient time has elapsed to permit reception of echoes from the most distant area under inspection, the transmitter sends out another pulse. For example, a radar with an interval of 2,000 μs between transmitted pulses would have a maximum range of 186 miles (Equation 1). Radar pulses vary from a few microseconds to several milliseconds in duration, with rates ranging from 1 to a few hundred hertz.

FIGURE 1



A pulsed radar system with targets (top) and the associated relative amplitudes and time intervals (bottom). Each short transmitted pulse is followed by a relatively long interval during which the receiver is active. The smaller amplitude echos (A) are from the smaller and less reflective tree, while the larger amplitude echos (B) are a reflection from the more distant but larger and more reflective building. The round-trip time from the transmitter (T) to each object can be used to calculate the distance of each object from the transmitter and from each other.

The strength and direction of a radar echo are functions of the size, distance, altitude, and composition of the object causing the reflection. In general, larger, closer, more reflective objects produce stronger echoes or signatures. The effect of an object's composition on its reflectivity depends on the frequency used. For instance, the absorptive surface coating used on United States and Russian stealth aircraft, while effective at microwave frequencies, has little effect in the 3 to 30-MHz range.[1]

Anyone who has listened to transmissions from the OSCAR satellites can appreciate the effect that relative motion has on a radio transmission. The Doppler effect describes this apparent increase in frequency of a signal transmitted from a rapidly approaching object, as well as the apparent decrease in frequency of a signal transmitted from a rapidly retreating object. The apparent shift in frequency is a function of *relative* velocity. It doesn't matter if the transmitter is stationary and the object is moving, or vice versa; the effect is the same. The greater the relative velocity, the greater the shift in frequency. From the following equation

$$\textit{Observed Frequency} = \frac{\textit{actual frequency}}{1 - (\textit{relative velocity/speed of light})} \qquad \textbf{(2)}$$

it's obvious that, for appreciable changes in the observed frequency, relative velocity must approach the speed of light.[2] The Doppler shift associated with speeds approaching 20 or 30 miles per hour is only on the order of 0.1 Hz. To the dismay of many motorists, Doppler radar technology has been perfected to the point where relative velocities of this magnitude can be determined to within a few miles per hour.

HF radar

Many Amateurs have the impression that HF radar (radar operating in the 3 to 30 MHz range) is a recent development. However, although the majority of radars currently in use in the United States operate at UHF and beyond, the earliest radar units relied on HF signals.[3] HF radars were used as early as the mid-1920s to measure the height of the ionosphere. The first operational military radar, the Chain Home system, installed by the British in 1938 to detect approaching bombers, operated at 25 to 30 MHz.[4] This line-of-sight radar system operated in the HF spectrum simply because reliable microwave transmitters of sufficient power didn't exist. It wasn't until World War II that effective microwave systems, duplexers, and other technologies required for VHF, UHF, and microwave radar systems were developed.

The first radars were good for determining range, but not direction. This was mainly due to the wide angle radiation beams provided by relatively small HF antenna systems. By decreasing the wavelength, it was possible to build antennas with narrow rotatable beams. These smaller and lighter narrow beam antennas were also more impervious to jamming because they were relatively insensitive to signals other than those arriving head on. During World War II, the German HF radar system was easily jammed because of the wide beams of their receiving antennas. In comparison, the United States Aircraft Warning Radar system used 100 to 200-MHz signals, and Japanese bombers used radar operating at 200 MHz. These VHF systems, with their narrow beam antennas, were seldom jammed.

Compared with early HF radar systems, today's UHF and microwave radars have several advantages: better support for weather observation; better low angle coverage; wider bandwidth for fewer problems with interference; narrow antenna beamwidth for improved resolution; and smaller, lighter systems that can be easily mounted on aircraft, tanks, and ships.[4] However, there continues to be a great deal of interest in HF radars — even those that operate as line-of-sight devices. HF systems are generally less expensive to build, have less critical antenna construction requirements, and suffer less interference from rain and insects than millimeter-wave radars.

Conventional radars that use moderate power microwave signals are line-of-sight devices, which means that their maximum operating range is limited by the curvature of the earth. Although airborne radars can be used to monitor the oceans and remote locations, 24-hour operation of an airborne system is expensive and may be impossible in inclement weather. There is currently a great deal of interest in OTH radar that uses high powered signals in the HF band, because these signals can propagate beyond the horizon and reflect from approaching aircraft and missiles.

Over-the-horizon radar

It's somewhat ironic that the very problems which hampered the widespread use of HF radar systems in the late 1930s — namely interference from skywave propagated, long range backscatter echoes from the earth's surface — has lead to a renewed interest in HF systems. Today, armed with Doppler radar techniques and digital computer signal processing, the propagation and interference problems of the past can be addressed.

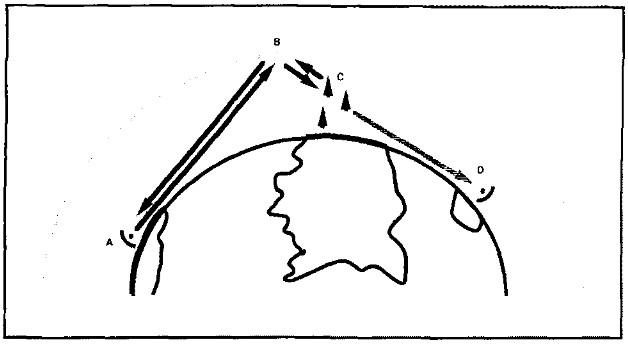
Capabilities

OTH systems provide a number of advantages over conventional radar The most obvious is the great surveillance range they support, from 500 to over 5,000 miles. Also, unlike conventional radar, OTH can detect aircraft at any altitude between the earth's surface and the ionosphere.[5] OTH waves are also relatively undeterred by passive stealth techniques that operate at microwave frequencies, like smooth shaping and the coating of aircraft with RF absorbing materials.

In addition to their absolute range, OTH systems are rated for their ability to resolve target range, angle, and velocity. For a single hop backscatter system, the range resolution is typically from 1-1/4 to 25 miles. Angular resolution, largely a function of antenna beamwidth, is typically 30 miles at a range of 1,900 miles. Target velocity resolution, a function of the operating frequency, is typically about 16 miles per hour at 25 MHz.[6]

Although OTH is best suited for aircraft detection, it can also be designed to detect ships and ballistic missiles, or to provide information about wind and wave conditions over the oceans.[5] An OTH system design is very target specific; it must be designed for the detection of either aircraft, ships, or missiles. The type of waveforms generated by the transmitter and the type of signal processing employed in the receiver system are highly target dependent.

FIGURE 2



OTH radar systems can be designed to operate in either forward or backscatter modes. In backscatter, pulses from the transmitter site (A) are reflected from the ionosphere (B). The earth's surface and targets, like missiles (C), reflect the radar pulses to the ionosphere (B) and to the transmitter site (A). In the forward scatter systems, pulses from the transmitter (A) are also reflected from the ionosphere (B). However, in this system, the receiving antenna is located distal to the area of interest(D).

Operation

Like conventional radar, OTH systems use pulsed RF to illuminate a target area. A major difference between conventional and OTH systems — at least in backscatter systems — is that the ionosphere affects both the transmitted and reflected signals (see Figure 2). OTH receiving systems rely heavily on digital signal processing technology. The composition of reflected radar signals, influenced by the surface of the earth and other objects that reflect HF signals, must be preprocessed by high speed digital computer systems. Only then is the data suitable for display and inspection on a video terminal.

Because OTH illuminates such a large portion of the earth's surface, land and sea echo (clutter) greatly overshadows the returned signal from targets. The dominant source of clutter is the earth's surface. Irregularities on the land and sea, trees, undergrowth, rocky outcrops, buildings, power lines, and towers all contribute to clutter. Natural phenomena, like meteor trails and auroral and equatorial field irregularities, as well as ionospheric instability, add to the clutter.[1] Luckily, Doppler techniques can be used to minimize clutter because the earth's surface is stationary, or at least moves slowly in relation to aircraft and missiles.

OTH radars are Doppler radars. They extract the desired signals from the background noise on the basis of the Doppler shift of signals reflected from moving targets. Signal detection becomes more challenging when the target and background have similar speeds, like ships moving at the same velocity as the sea.[7] With Doppler processing, the transmitter beam must illuminate the target area for a period (dwell time) sufficient to distinguish moving targets from the background clutter. Depending on the clutter and the nature of the target, the dwell time can range from less than one to over ten seconds.

As in Amateur HF communication, the ionosphere determines the range of operation for OTH systems. Although most systems can use any segment of the HF spectrum, they usually operate near Maximum Usable Frequency (MUF) to minimize propagation losses. The MUF, the highest frequency that can support propagation that relies on the ionosphere, varies with the state of the ionosphere, time of day, season, and amount of sunspot activity. Ionospheric conditions are usually monitored in real time through vertical sounding of the ionosphere, oblique sounding, and, in a backscatter OTH system, by observing the signals which have been reflected by the earth.

In addition to the problems associated with propagation losses, the ionosphere places additional demands on OTH radar systems. For example, there are usually multiple paths from the radar transmitter to target, as well as from the target to the receiver. In addition, the velocity of propagation is frequency dependent. Together these factors effectively limit the range resolution of an OTH system because short pulses can be distorted to the point of being useless. The refraction of signals by the ionosphere is frequency dependent, allowing a specific area to be illuminated by a narrow range of frequencies. For example, a signal on 10 meters will normally illuminate the earth's surface at a much greater range than a signal on 80 meters. Finally, the ionosphere is in constant flux; propagation characteristics can change unpredictably from moment to moment. An OTH system must be robust enough to handle losses due to polarization mismatch, ground reflection, and the focusing or defocusing of the radar signal due to the ionosphere.

Transmitter and receiver systems

OTH transmitters must provide not only very high power output (on the order of a megawatt or more) but operate over a wide range of frequencies. They must also have the ability to change operating frequency instantly (frequency agility) and produce highly accurate and stable carrier frequencies (for Doppler detection) with high spectral fidelity.[8] Non-Soviet OTH systems use several transmitters operating simultaneously to meet these requirements.

The transmitted signal can be continuous wave (CW), simple pulse, frequency modulated-continuous wave (FM-CW), chirped pulse, or some other coded waveform.[3] Regardless of the type of modulation employed, it's usually desirable to shape the transmitted pulse to minimize the spectral energy contained at frequencies far removed from the carrier (similar to the key click filter in your CW transmitter).[3]

The main operating variables in the transmitter system are the output frequency, signal bandwidth, pulse repetition rate, and pulse duration. Wide area coverage requires a wide range of frequencies. The radar scanning area (footprint) is moved when the frequency is changed from the low to the high end of the HF spectrum, with higher frequencies used to scan the more distant target areas. Because dwell time must also be considered at each frequency, coverage of a large area can take considerable time. Scan time can be reduced by transmitting and receiving simultaneously at several frequencies, or by transmitting a very wide beam and employing several narrow beam receiving stations.

Good range resolution requires a wide signal bandwidth. A 100-kHz bandwidth, about the maximum that the ionosphere can support at any given instant, corresponds to a range resolution of a little less than a mile.[3] The use of wide bandwidth transmitter pulses has the added advantage of reducing echoes from natural targets.[3]

A low pulse repetition rate avoids range ambiguities. At 50 Hz, for example, the unambiguous range is about 1,900 miles. However, at low pulse repetition rates, Doppler signal

processing becomes more difficult. The pulse repetition rate is, therefore, a compromise between range and velocity resolution. In practice, OTH pulse repetition rates range from several to tens of hertz.[8] A long pulse duration is required for long range detection. Typical pulse duration varies from hundreds of microseconds to several milliseconds.[8]

In addition to matching the general abilities of transmitter systems — including frequency coverage, frequency agility, and frequency stability — the broadband receivers used in OTH systems must operate in high levels of active interference (from radio transmitters and meteors) and from passive interference (ground and ionospheric reflections). The most sophisticated components of the receiver system, however, don't deal with signal capture but rather with signal interpretation. A single receiver site may require a network of a dozen or more large computer systems, all dedicated to processing the received signals.

Forward scatter and backscatter

OTH radars can operate with either backscattered waves, where a reflected signal is received at or near the transmitter site, or with forward scattering, where the incident and scattered waves propagate in the same direction (see **Figure 2**).[6] Backscatter OTH systems can detect the presence of a target and measure its location. As in conventional Doppler systems, the transmit/receive time delay determines range. Highly directive antenna systems determine angular coordinates.[6]

Backscatter systems require very high power transmitters (in the megawatt range) for reliable operation, because the losses associated with the ionosphere affect not only the transmitted signals, but the reflected waves as well.

Forward scattering systems can achieve very long operating ranges, on the order of 5,000 miles — about twice the range of backscatter systems. Much lower power transmitters can be used with forward scatter than with backscatter, which is an advantage. Disadvantages include the inability to measure range. Only the presence or absence of a target can be detected. This obstacle can be overcome by using multiple receiving sites, but this is usually prohibitively expensive and complex.

Another constraint associated with forward scatter systems is that the transmitter and receiver sites must be located on either side of the area of interest. For example, to monitor the Soviet Union, the United States installed forward scattering transmitters in Asia, the Pacific, Japan, Taiwan, and the Philippines and receivers in West Germany, Italy, and the United Kingdom.[6] An additional problem associated with forward scattering is that of maintaining synchronization of transmitter and receiver systems. High stability oscillators in the receiver system that make use of periodic synchronizing signals from the transmitter have been used to help maintain synchronization.

Antenna systems

Antennas for OTH radar must fulfill an almost impossible list of requirements. They must they have high directivity and the ability to scan the entire azimuth rapidly; they must also be steerable in elevation. OTH antenna systems must produce high gain, on the order of 20 to 30 dB, to provide both high signal-to-noise and clutter-to-noise ratios. (In this context, noise is considered to be due to lightning and broadcast transmitters and clutter is considered to be echoes from non-target objects.) Although fulfilling these requirements at a single frequency would be challenging enough, OTH antenna systems must meet the aforementioned specifications for the entire HF band.

Perhaps the most awe-inspiring feature of OTH antenna systems is their sheer size. Systems extending thousands of feet in length and over a hundred feet high are not uncommon. Also, OTH antenna systems often use separate transmit and receive antennas. Although duplexing (using the same antenna for transmitting and receiving) is common in millimeter-wave radars, it's seldom used in OTH systems. Because of the high transmitter power, sometimes on the order of several megawatts, the transmitter and receiving antennas are commonly separated by many miles. By its very design, forward scatter OTH necessitates the use of separate transmit and receive antennas. The high power associated with the transmitters places a number of demands on the construction of the transmitter antenna — including the need for massive feedlines and antenna elements. On the other hand, receiving antennas can be constructed with conventional materials.

Both horizontal and vertical polarization can be used with OTH systems. When vertical polarization is employed, a metallic ground plane is usually placed in front of the antenna to reduce losses — especially if the antenna installation is located inland.[8] The ground plane for OTH systems most often takes the form of a wire screen extending 150 or more wavelengths in front of the antenna.[3] Above ground screens are used to minimize losses caused by snow in places where there's snow on the ground for much of the year.

Installations

There are probably several dozen OTH systems in operation around the globe at any one time. Although there's almost no mention of many of these systems in the literature, several have been described, if only by observers. Perhaps the most infamous OTH system is the Russian Woodpecker (so named because of the characteristic pulse rate of about 10 Hz). This extremely powerful system, located in the Ukraine, was first widely heard during the winter of 1976-77.[9] Recently, a missile-tracking OTH system based in Siberia has been reported.[10] This system uses a much higher pulse repetition rate than the woodpecker. Its sound has been likened to that of a bumblebee.

Although we tend to attribute all OTH signals to Soviet systems, there are a variety of sources for these characteristic pulses — including several sites in the United States. A system located in Caribou, Maine, for example, uses 21 100-kW CW-FM transmitters, of which seven may be transmitting simultaneously.[9] The dipole transmitter array, separated from the receive antenna by 100 miles to minimize interaction, is 130 feet high and 2,225 feet wide. The operating range for this backscatter system is from 500 to 1,800 miles.[6]*

**Wonder how my 2-watt handheld would work with one of these antennas? Ed.*

The Wide Aperture Research Facility (WARF) in Los Banos, California is used for monitoring ocean conditions and for tracking ships in the Pacific and Caribbean. This FM-CW radar system, operating from 3 to 30 MHz, uses separate transmit and receive antennas located 115 miles apart. The receiving antenna, an array of 256 twin whip doublets that extends 1.6 miles, has a beamwidth of only 0.5 degree at 15 MHz.[7]

OTH development isn't limited to the United States and Soviet Union. Australia operates Jindalee, an OTH system near Alice Springs (near the center of the country), to monitor its northern coast.[5] The transmitter site supports 16 CW transmitters that drive 16 log-periodic antennas. The receiving antenna consists of 492 elements, with an aperture of about 900 feet.[7] An extensive ground screen is used to maximize transmitter efficiency. The Hall Beach system, located in Melville, Canada produces a 3-MW signal with a log-periodic antenna.[8]

Apparently, each branch of our armed forces conducts its own OTH research. The United States Air Force, which already maintains OTH surveillance of both East and West Coasts, will soon begin transmitting OTH signals from Hanscom Air Force Base in Massachusetts.[11] The Navy's MADRE system was developed at the United States Naval Research Laboratory on the Chesapeake Bay in the early 1960s. This 5 to 50-kW system uses a phased array 320 feet wide by 140 feet high. The Table Mesa OTH at Boulder, Colorado, first put into use in the mid-1950s, has reportedly been used to scan the Gulf Coast.[7]

The Pentagon is building a new OTH backscatter network to cover the coastline of the United States. The first unit in this network, built in Maine, ran its first tests last year. Each radar site is capable of tracking targets from 500 to 1,800 miles away and produces 1 MW of FM-CW pulsed radar at angles of 6 to 30 degrees. The Maine unit covers the area from the East Coast of the United States to Cuba, to Greenland, and to Iceland, with an operating frequency range from 5 to 28 MHz. The receiving antenna is 5,000 feet long and uses 28 VAX computers for signal and data processing. An even more elaborate West Coast unit has an 8,000 foot long receiving antenna.[1]

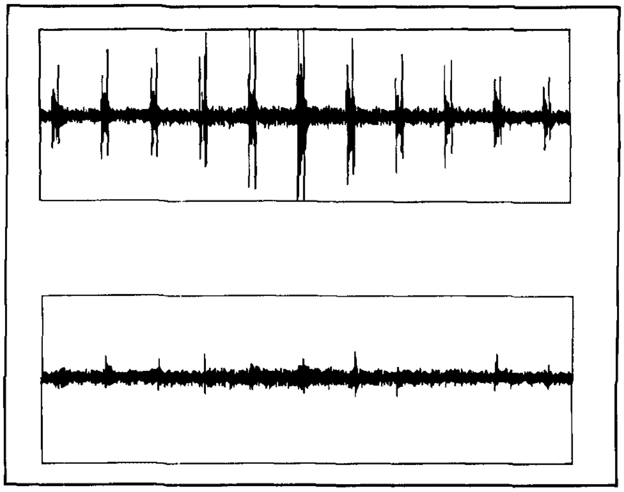
Coping with OTH radar noise

Using a highly directive antenna system will minimize the effects of OTH, especially when the radar signal is broadside to the antenna's major lobe. However, because most of us don't have six-element Yagis on every HF band, we need another more general solution. Noise blankers, while not a panacea, can help minimize the effect of OTH QRM.

Most modern transceivers come equipped with noise blankers designed to reduce intermittent noises. Unlike noise limiters — which work by limiting the amplitude of noise spikes to the average signal intensity — noise blankers actually interrupt the IF signal during the exact time of each noise spike. Short duration spikes are handled easily by most noise blankers, allowing the desired signals to stand out clearly from the background noise (the instantaneous breaks in the desired signal aren't usually noticeable).

Unfortunately, the effectiveness of noise blankers diminishes as the noise increases in amplitude and duration. Longer duration high amplitude noise, like the woodpecker noise associated with OTH radar (**Figure 3**), is handled poorly by simple on/off noise blankers. More capable noise blankers, with adjustments for duration and blanking level, are best for coping with OTH noise. However, even with the best noise blankers, OTH signals can't be eliminated completely (**Figure 3**). When using a noise blanker on OTH signals, work with the lowest blanking level possible. Using a higher blanking level than necessary will only add distortion to the desired signal.

FIGURE 3



Recording of an OTH signal made on February 10, 1989 at 20:40 UTC (top). At the time of the recording, the signal, which had a pulse rate of approximately 10 Hz, could be heard from 10.495 to 10.509 MHz. Notice the cyclical nature of the signal's peak amplitude, implying that the carrier frequency was systematically changed from pulse to pulse (FM-CW modulation). The illustration at the bottom is a recording of the same signal made at 20:41 UTC. For this recording, the receiver's noise blanker was adjusted to minimize the OTH signal. Note that, although a high quality adjustable noise blanker was used, a considerable amount of the unwanted signal remains.

Summary

OTH radar, and its characteristic interference, is here to stay. The situation on the HF Amateur bands is likely to become much worse when the new high power systems become active on both east and west coasts of the United States. Because these systems apparently have free reign over the entire HF spectrum, our most immediate recourse as Amateurs is to develop more effective receiver circuitry for rejecting their signals. A long term solution will no doubt require international legislation to limit OTH systems to non-Amateur frequencies.

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PRODUCT REVIEW

Smith Design 107 Spectrum Probe

Ever long to own a spectrum analyzer? As a project builder and radio repairer, I sure have! Analyzers are invaluable for a wide range of design and service tasks, and most professional labs have them. Yet the price tags on these electronic marvels have never been in keeping with my "Amateur" status.

Fortunately for those of us on a budget, Smith Design now offers a new device called the 107 Spectrum Probe. For about the price of a 2-meter FM rig, you can team this compact device with any inexpensive oscilloscope and perform many important tests once reserved for expensive and cumbersome laboratory setups.

In terms of technical specifications, the probe's frequency coverage spans from below 1 MHz to more than 100 MHz with a dynamic range of 50 dB or greater. Sensitivity is around 100 μV, and frequency response is specified at $\pm$2 dB from 5 to 100 MHz. IF bandwidth is 180 kH at -3 dB, providing an ultimate resolution of about 0.5 MHz. Maximum CW input is +15 dBm or 1 volt at 100 MHz, becoming progressively greater at lower frequencies (the probe has a 10-pF input isolation capacitor). Sweep rate is 6 ms/100 MHz, and horizontal linearity is $\pm$10 percent. The probe measures 7.5" × 1" × 1" and weighs just 2 ounces. External power is supplied by a small wall transformer.

I found the 107 very easy to set up and operate. The probe has no user controls — all adjustments are made at the scope. The 5'6" probe cable has two connectors. There's a line jack for the wall transformer and an RCA plug with BNC adapter for vertical input. Horizontal timing is provided by a sync pulse on the probe's composite waveform.

To obtain a spectrum display, I set vertical gain for 50 mV/div (DC coupled) and triggering for 0.5 ms/div, negative slope. I then repositioned the baseline to the bottom of the screen, pushing the horizontal sync pulse out of view. Finally, I checked the trigger level for stability. "Zero" MHz was marked on the left of the display by a strong carrier (an internal local oscillator). Each division across the screen represented approximately 10 MHz, with the last corresponding to 100 MHz. Because my scope has x10 sweep expansion, I could zoom in for a closer look at any 10-MHz segment over the probe's 100-MHz range. To check sensitivity, I touched the probe to my finger and brought up several broadcast carriers on the display

For my first test, I connected a homebrew 20-meter QRP rig to a dummy load and keyed it. I then placed the probe tip near the load, positioning it so the 14-MHz fundamental measured 50 dB, or five divisions, above the baseline (suggested zero reference). At this level, the stronger mixer spurs and a harmonic began to poke up through the noise floor, allowing me to estimate frequency and amplitude For spotting exact frequencies, I used a signal generator as a tunable marker. This "quickie" sampling procedure yielded nearly identical results to those obtained in a lab evaluation of the same radio.

A lot of the "business" that goes on in RF circuitry is difficult to observe without some form of spectrum analysis. Consequently, I have come to rely on the probe a great deal. Being able to "see" what occurs when two signals mix, when a stage is overdriven, or when a filter is mistuned makes debugging and optimizing much easier. Tough jobs, like checking a mixer-driven transmitter strip for FCC compliance, are simple fare. I've also found the 107 very "test friendly," in that it rarely loads down circuitry and often picks up usable signal levels without physical connection. However, because of this sensitivity, the manual recommends using caution around power amplifiers to avoid burning out the probe's input amplifier. I also observed some desensing in high RF fields (the probe case is unshielded).

In addition to functioning as a Hi-Z voltage probe or looped current probe, the 107 interfaces readily with standardized systems. A slip-on adapter converts the probe tip to a BNC-compatible plug. By adding a "T" connector and terminating load, you can plug directly into calibrated generators, attenuators, mixers, and RF bridges. When terminated, the probe is virtually transparent throughout its range.

Although this is a great product, it's important to remember that the 107 isn't a direct replacement for a $10,000 lab instrument. Dynamic range is 50 dB versus 100, accuracy is $\pm$3 dB versus 0.3, and coverage is 0 to 100 MHz versus 0 to 1 GHz. But if some tasks are beyond the gambit of the 107, it handles a host of others economically and well. For example, it's a natural for EMI and RFI detection, service work, digital and HF analog design, classroom and laboratory training, production control, and many other real-world applications. And, for the Amateur who builds or repairs ham gear, the Spectrum Probe is clearly one of the most useful tools I've seen. I recommend it.

Smith Design 107 Spectrum Probes are available directly from Smith Design, 1324 Harris Road, Dresher, Pennsylvania 19025. Credit card orders are accepted and may be placed by calling (215)643-6340. The list price is $249, including adapters, power supply, and storage case. Significant discounts are available — including a special $199 direct mail price to Radio Amateurs.

de K1BQT

Circle #120 on Reader Service Card.

The Complete DX'er

Amateur Radio has several very good authors. One of the best is Bob Locher, W9KNI. Bob's ability to spin an interesting yarn about a relatively technical subject is unique. Many of you will remember Bob from his series of articles in *Ham Radio HORIZONS*, the "Dxer's Diary." Reading the "DXer's Diary," you shared the joys of working rare DX and the misery of having missed out.

Considerably revised and updated, this new edition of an all-time classic is sure to be a best seller. Bob assumes that his reader wants to work DX. He then sets out to weave a carefully crafted, yet eminently readable, tale of how to do so. From calling CQ to spending hours listening to static crashes on the low bands, Bob gives you the benefit of his years of operating experience and DXCC Honor Roll status.

Special coverage, including operating hints and strategies, is given to working DX in contests — often one of an operator's best opportunities to work new countries. For instance, a good single operator station can work over 100 countries in CQ's World Wide™ contest.

Locher also gives special treatment to working DX through pileups. Read *The Complete DX'er* and you'll have all the tricks of the trade from one of the best at your fingertips. You'll find it's better to listen to a pileup with the rare DX station already in your log than to spend hours calling frantically, and sometimes unsuccessfully!

This book is great reading for the experienced operator and "must" reading for the newcomer. You'll learn how W9KNI has been so successful in his years chasing DX. Who knows, maybe you'll join him on the DXCC Honor Roll someday soon.

The Complete DX'er is available from the **HAM RADIO** Bookstore for $11.95 plus $3.75 shipping and handling.

de NX1G

THE HANDY ANTE SWITCH

Frank W. Smith, W4EIN, 2023 Haven Crest Drive, Chattanooga, Tennessee 37421

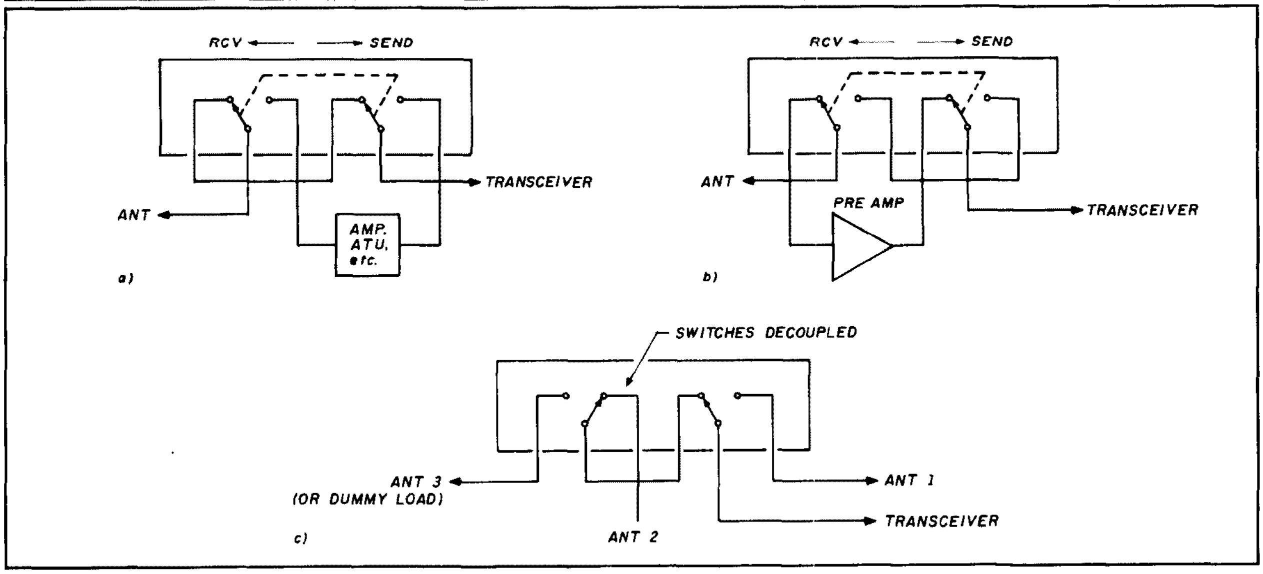
Several years ago I needed a double-pole, double-throw (DPDT) coax switch in my station. I couldn't find a commercial supplier for such a device, so I coupled two Radio Shack CB-type antenna selector switches mechanically. Basically, they were single-pole, double-throw coax switches. These worked quite well for a number of years — serving primarily to insert a homebrew Class-C amplifier into the antenna feedline when in the send position and to bypass the amplifier in the receive position.

Uses

I discovered that this switching arrangement lent itself to a number of applications. You can insert a preamplifier into a transceiver input for receiving and bypass it on transmit, bypass the antenna tuning unit on designated bands, and select an antenna or dummy antenna. These applications are shown in **Figure 1**. The switch is also useful for making immediate real-time comparisons under operating conditions — like reception with and without a preamp or effectiveness of a power amplifier with a contacted station.

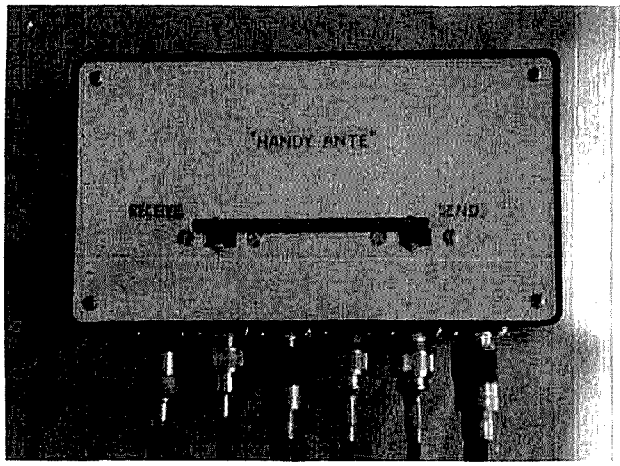
Eventually, wear and tear got the best of the rather puny contacts on the CB switches, and I needed to replace them. But alas, the Radio Shack switches were no longer available, and I couldn't locate any other similar devices. So, I decided to concoct my own. The result was sufficiently successful to warrant passing it on.

FIGURE 1



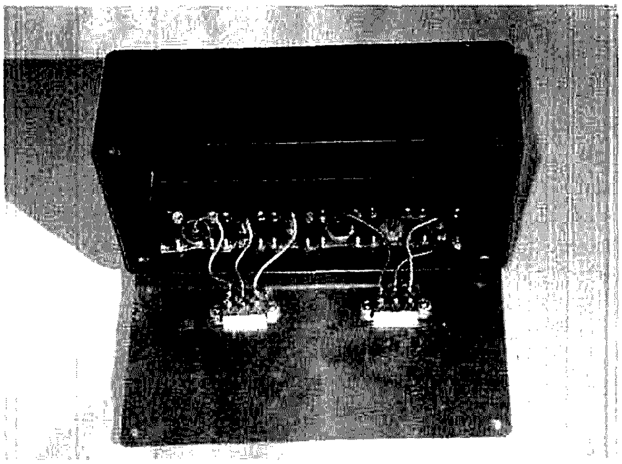
Some typical applications for "Handy Ante."

PHOTO A



The DPDT coax selector switch (dubbed the "Handy Ante") in operating position.

PHOTO B



"Handy Ante" with front panel open. Note lug in lower right corner for grounding panel.

PARTS LIST

Quantity	Description
1	*Housing, Radio Shack hobby box no. 270-232, or similar*
2	*Slide switches, DPDT, 125 volts AC, 3 A, Radio Shack no. 275-403*
6	*Coaxial cable receptacles, chassis mount, type SO-239, Radio Shack no. 278-201*

Description and construction

Essentially, the device consists of two slide switches coupled together mechanically and terminated electrically in six SO-239 coaxial cable receptacles. The construction is quite simple and is shown in Photos A and B. I built it into a Radio Shack 7-3/4 by 4-3/8 by 2-3/8 inch plastic experimenter's box with an aluminum panel. This box is a little large for this application, but it does permit in-line mounting of the six coax receptacles. It's critical to bond the SO-239 receptacles together electrically; I used copper braid from a scrap of coax. The leads from switches to receptacles should be short.

Switches

The slide switches are double-pole, double-throw, but the poles are paralleled to increase the current capacity. The switches are rated 3 A at 120 volts. Paralleling the contacts should make them good for 6 A, but because they are intended for use in low frequency power circuits, I'd hesitate to give them this rating at RF. I can say from experience that they will handle the output of a 200-watt amplifier and suspect that they will do considerably better.

The switches are coupled by a length of wooden dowel. I used one 1/4 inch in diameter and 3-3/4 inches long. Drill a 5/64-inch hole in each of the plastic switch levers; then mount the switches and carefully mark and drill the dowel to match. I mounted the dowel with no. 2-56 machine screws and hex nuts.

Be sure that both switches are in the same position when you mark the dowel for drilling. If you take reasonable care, the switches will move smoothly and positively from one position to the other, with the dowel used as a handle. I used no. 18 stranded hookup wire to connect the switch. Make sure the leads are short and have sufficient current capacity. If you use a metal panel, connect it to the coax receptacles.

Mount the completed unit solidly to a wall, shelf, or other stable mooring near your operating position. Now all you need to put the "Handy Ante" to work are a few coax jumpers. The coax you choose for these jumpers should have the same characteristic impedance as that of the original transmission line.

Performance

The switch in the circuit at my location seems to have minimal effect on VSWR or general performance of the antenna system. I found there was an increase in VSWR from 1:1 to 1.5:1 at 28 MHz when I inserted the switch in the line. This isn't a major problem and was the maximum change noted between 3.5 MHz and 28 MHz.

A note of caution

If you use the switch to insert a preamplifier or some similar device used mainly for receiving, make sure the device has protection in case you forget to throw the switch to the transmit position to send. I use back-to-back diodes mounted directly across the preamp input terminal from the transceiver. This crowbar-type protection causes the transceiver to transmit into a virtual short circuit and trip off because of its protective system. Other types of protection may be best for different setups.

In closing

You might ask, "Why, in these days of sophistication and automation, should I resort to a manual send/receive switch?" The system is simple and reliable, and lends itself well to on-the-spot comparative testing. The switch also operates easily and quickly. Its only drawback is operation on break-in modes. I use "Handy Ante" on a routine basis to insert my old Class-C amplifier in the feedline when needed. I also use it to insert a 10 or 15-meter homebrew preamp for receiving, combining functions A and B of Figure 1.

Microwaves

Bob Atkins, KA1GT

Starting something new is always a challenge. I thought I'd begin this month by introducing myself and talking about the possibilities for this column. As some of you may know, I was the author of *QST*'s column on microwave technology from 1980 to 1989. When *QST* decided not to run a regular monthly microwave column late last year, *Ham Radio* graciously offered me the opportunity to join them and continue coverage of the bands above 1 GHz.

I have some ideas about how such coverage might be presented, but I would be interested in what you would like to see. My current idea is to present a mix of operating news, technical presentations, and simple construction projects. These would be aimed mostly at the microwave bands, but would not exclude coverage of topics also related to operation in the VHF and UHF parts of the spectrum. Your input to such a column is *vitally* important if it is to present information concerning state-of the art developments in microwave equipment and operation. If you have short news items, technical hints, or microwave-related questions, I urge you to send them along to me at the address at the foot of this column. There is a minimum two month lead time between my writing a column and its appearance in *Ham Radio*, so prompt reporting of activity news is especially important.

Ham Radio is always interested in publishing longer technical articles in the main body of the magazine, so if you have such technical material, by all means send it directly to the magazine at the address on the contents page.

It's difficult, in writing a column of this type, to determine the level of knowlege I should assume on the part of the reader. As much as possible, I will aim at general Amateur interest. There are two reasons for this. First, I want to encourage readership, nothing turns off readers like a page full of equations! Second, I'm not sure I can write at a level more complex than that! Plans for upcoming columns include coverage of optical communications systems including laser transmitters, receivers, and propagation, and articles on the various modes of microwave tropospheric propagation — from line of sight through tropospheric scattering to ducting. Along with these I hope to have some news items. Once again, that depends on reader input. So please write, and between us we'll see what we can do to maintain coverage of the microwave scene.

Please send your questions and comments to me at 103 Division Avenue, Millington, New Jersey 07946.

Meet Bob Atkins

In January's Editor's Notes I promised all of you a new column for microwave and VHF/UHF enthusiasts. We know that many of you have missed Joe Reisert's monthly missive "VHF/UHF World," so this month we are pleased to present our first installment of "Microwaves," written by Bob Atkins, KA1GT. Bob holds a Ph.D. in chemistry and currently works in the field of materials research at AT&T's Bell Laboratories. He was first licensed as G8EKB (a UK no code license) in or around 1969. He was active intermittently on 144, 432, and 1296 MHz until 1976, when he left England for Texas. In 1979 Bob moved to Connecticut, was licensed as KA1GT, and operated SSB/CW on 144, 432, and 1296 MHz and wideband FM on 10 GHz. From 1979 to 1981, he was also active on 432-MHz EME using a homebrew amplifier and antenna system. Bob made his first contacts on 10-GHz SSB/CW in 1981. Later that year he moved to New Jersey, where he can be found on the air from time to time on the VHF/UHF and microwave bands. He enjoys building equipment, and has recently completed construction of a fully relay-switched solid-state 10-GHz all mode transverter system for portable use. We hope you enjoy Bob's column and that you'll send him your comments and suggestions.

KA1STC

SWITCHABLE BANDWIDTH **CRYSTAL FILTER**

CW and SSB filter with one set of surplus crystals

By John Pivnichny, N2DCH, 3824 Pembrooke Lane, Vestal, New York 13850

You can design a filter for almost any bandwidth by selecting the proper coupling capacitors and termination resistances. Here's how to build a filter with two different bandwidths (CW and SSB) by switching only the capacitors. The termination resistance is held constant at 220 ohms and doesn't need to be switched.

Ladder-type filters

Simplified design equations for ladder-type filters where all crystals have the same frequency have been published in Amateur Radio literature.[1] Figure 1 shows a typical ladder filter. (See Reference 2 for more information.) The crystals are in series with the signal path and are "coupled" by capacitors with one end connected to ground. The filter is terminated with resistances at each end.

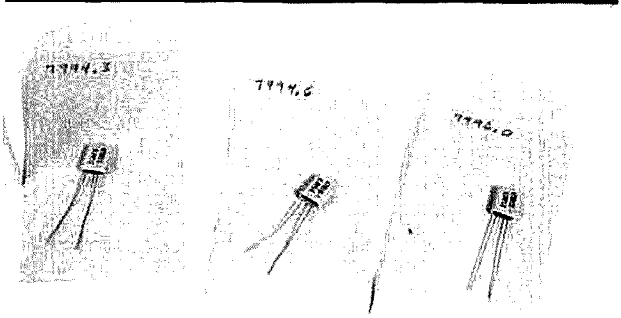
Crystals

My filter uses inexpensive surplus 8.0-MHz crystals in an HC-18 style package with wire leads (see Photo A.) I was able to get 50 of them through a swap and checked their frequencies before proceeding.

You can check your crystals by building the oscillator circuit shown in Figure 2 and Photo B. The terminal posts allow for easy connection and removal of each crystal's wire leads. I wrote frequencies to the nearest 100 Hz on a set of small envelopes and placed each crystal in its marked envelope directly after removing it from the test circuit. My envelope system prevents mixups and lets me select appropriate crystals for the filter.

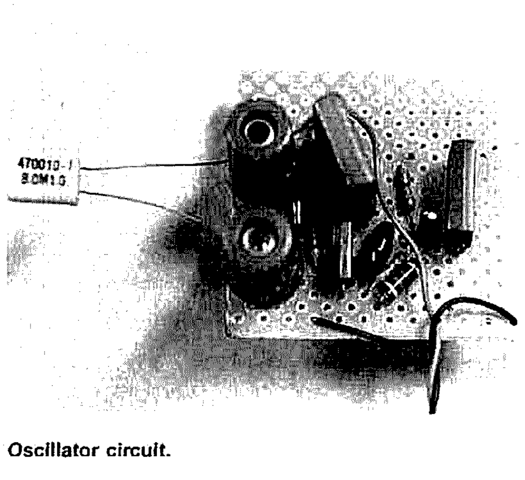
You'll need three crystals with frequencies within 200 Hz of each other. In most cases, it isn't necessary to start with 50 crystals. My crystals turned out to be very low Q, with frequencies ranging from 7995 to 8015 kHz. Nevertheless, I wound up with many sets of three within the required 200 Hz. The surplus crystals, although low in Q, make very satis-

PHOTO A



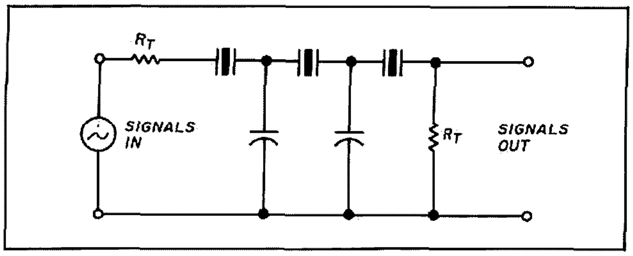
Surplus crystals.

PHOTO B



Oscillator circuit.

FIGURE 1



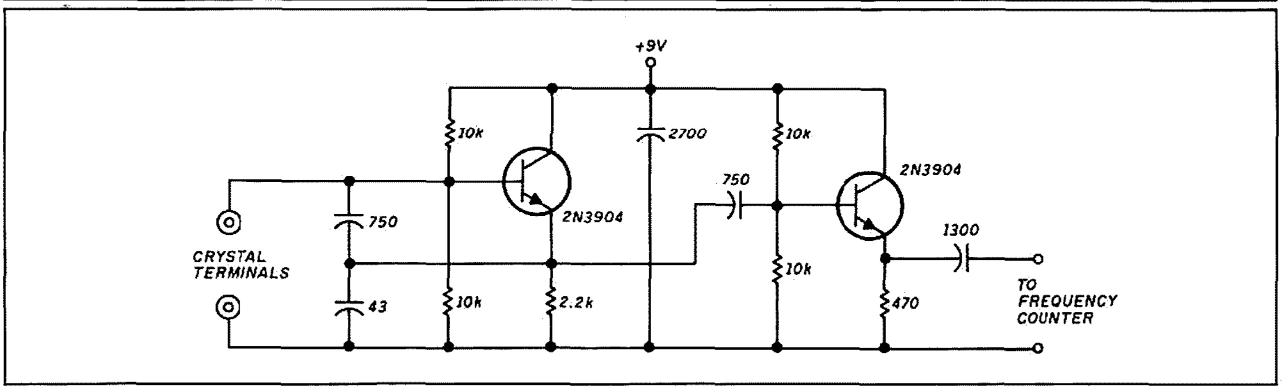
Typical ladder filter.

factory filters. The main difference you'll notice when using low-Q crystals is the higher insertion loss for narrower (CW) bandwidths. You can compensate for this loss in the IF amplifier that follows the filter.

Test circuit

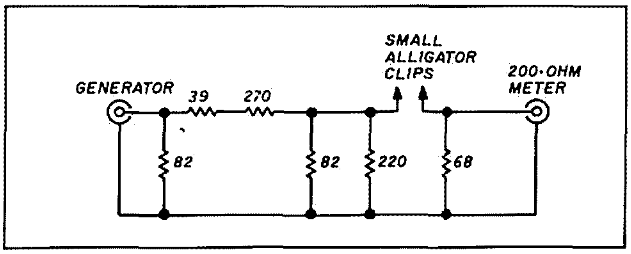
Once you've selected three crystals, you must measure their characteristics more carefully. You'll need a signal source whose frequency is adjustable ±20 kHz around the crystal frequency and a sensitive RF voltmeter. I used the circuit and hardware of my homebrew tuning dial[3] (adding 240 pF to the tank and adjusting the coil slug), which I tuned to 8 MHz for a signal source. My voltmeter is described in Reference 4. You'll also need a calibrated 3-dB attenuator[5] and the 50-ohm test circuit in **Figure 3**.

FIGURE 2



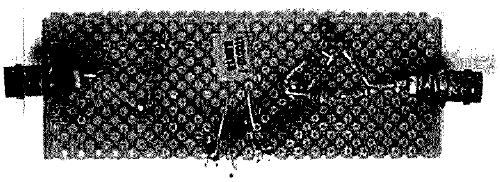
Oscillator circuit.

FIGURE 3



50-ohm test circuit.

PHOTO C



Test circuit.

You must have a 50-ohm system to measure the crystals because the design equations are based on 50-ohm measurements. The pi attenuator at the input of **Figure 3** guarantees a source between 50.05 to 51.82 ohms, regardless of your signal source impedance. At the output, I used the 200-ohm attenuator[5] in parallel with a 68-ohm resistor, followed by a 200-ohm voltmeter.

Following Hayward's procedure[1], insert each of the selected crystals into the test circuit shown in **Photo C** and tune the signal source for a peak reading on the voltmeter. This peak will occur about 1.8 kHz lower than the frequency measured with the oscillator circuit of **Figure 2**. Note the reading on the voltmeter. Now replace the crystal with low value fixed resistors (10 ohms is a good starting point) until you obtain the same voltmeter reading. This value will be the series loss (Rs) of the crystal. My surplus crystals measured 27 to 30 ohms; a high quality HC-6/U type will measure 5 to 6 ohms. Record the Rs measurement.

Replace the crystal in the test circuit, checking to make sure the signal source is still set for a peak reading on the voltmeter. Switch in 3-dB attenuation, note the new meter reading, and then switch back to 0 dB. Now tune the signal source above and below the peak frequency until you reach the same new meter reading. Record these two frequencies.

The difference between the two frequencies will be the bandwidth Δf (in hertz) of the crystal at the 3-dB points. My surplus 8-MHz crystals measured 1230 Hz. Hayward reported 294 Hz for his 3.5-MHz crystals and 96 Hz for high quality 3.5-MHz crystals.

SSB filter calculations

Design the SSB bandwidth filter (2.5 kHz) first, using the equations which follow. I took them from Hayward's article.[1] The circuit shown in **Figure 4** is a Butterworth three pole, tuned with series capacitors in the end loops. All capacitors and resistors have the same value.

$$C = 1326 \left[\frac{\Delta f}{B \times 0.707 \times Fo}\right] - 10\ (pF) \qquad \textbf{(1)}$$

$$Rend = \left[\frac{120 \times B}{\Delta f}\right] - Rs\ (ohms) \qquad \textbf{(2)}$$

where B = desired filter bandwidth in Hz
Fo = crystal center frequency in MHz
Δf = bandwidth measured in test circuit in Hz
Rs = crystal series resistance measured in test circuit in ohms

Using my crystals with Δf = 1230 Hz, Fo = 8 MHz. The calculated values are C = 105 pF and Rend = 217 ohms.

You can calculate the crystal's equivalent series circuit elements from **Equations 3** and **4**. A holder shunt capacitance of 5 pF is assumed. (See **Figure 5**.)

$$Cm = 1.326 \times 10^{-15} \left[\frac{\Delta f}{Fo^2}\right] (farads) \qquad \textbf{(3)}$$

$$Lm = \frac{19.1}{\Delta f}\ (henries) \qquad \textbf{(4)}$$

where:
Cm and Lm are the equivalent series capacitance and inductance, respectively.

I used the equivalent circuit to compute the frequency response of the filters in this article aided by a PC circuit analysis program.[6] With 8 MHz as Fo, I calculated Cm = 0.0255 pF and Lm = 0.01552 H. The computed and measured response of crystal no. 3 in the 50-ohm test circuit is shown in **Figure 6**. I also ran the analysis program for the SSB filter with and without the end loop tuning capacitors. These results are shown in **Figure 7**. It appears that the tuning capacitors eliminate the dip in the center of the passband. They also shift the overall response 500 Hz higher on the low frequency side and about 250 Hz on the high frequency skirt. This narrows the bandwidth slightly.

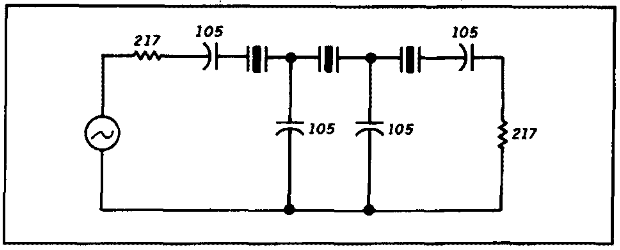
CW filter calculations

A much narrower bandwidth is desirable for CW. I chose 500 Hz, and using **Equations 1** and **2**, calculated C = 565 pF; Rend = 21.8 ohms. Note that Hayward[1] recommends that the crystal unloaded Q (Qu) exceed the filter Q by a factor of 10.

$$Qu = \frac{1.2 \times 10^8 \times Fo}{\Delta f \times Rs} \qquad \textbf{(5)}$$

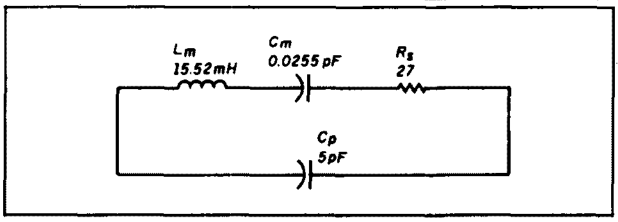
For my crystals, this was Qu = 28,900. A 500-Hz bandwidth at 8 MHz represents a filter of Q = 16,000, so a factor of

FIGURE 4



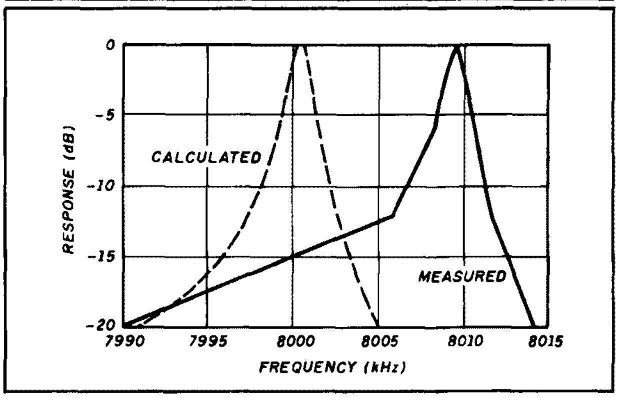
2.5-kHz SSB filter.

FIGURE 5



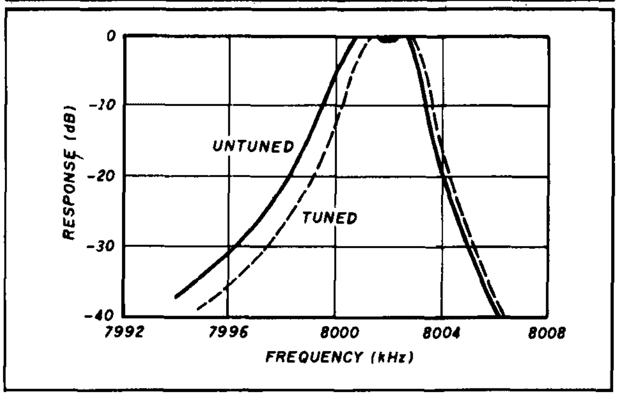
Crystal equivalent circuit.

FIGURE 6



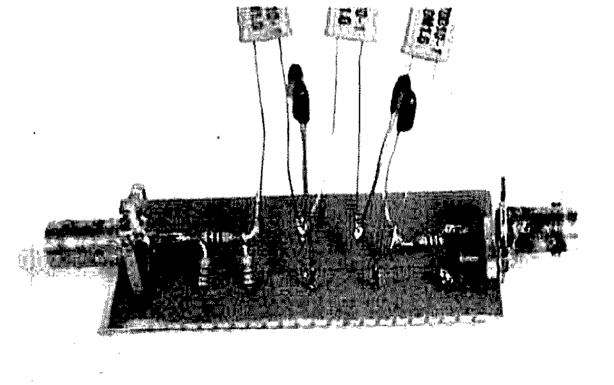
Test circuit response.

FIGURE 7



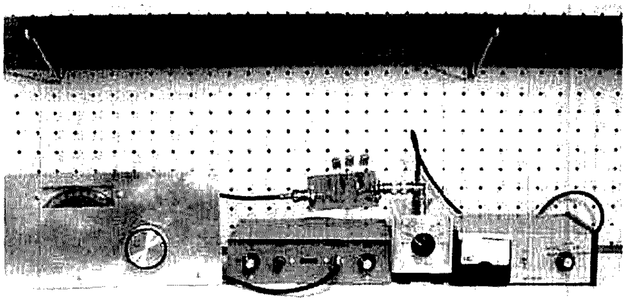
SSB filter, calculated data.

PHOTO D



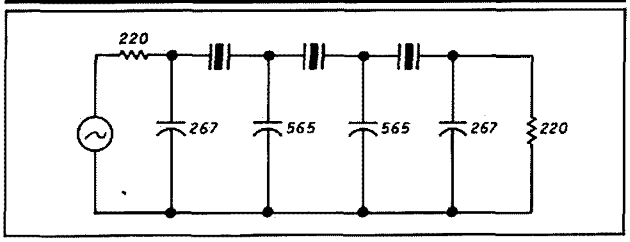
Trial filter.

PHOTO E



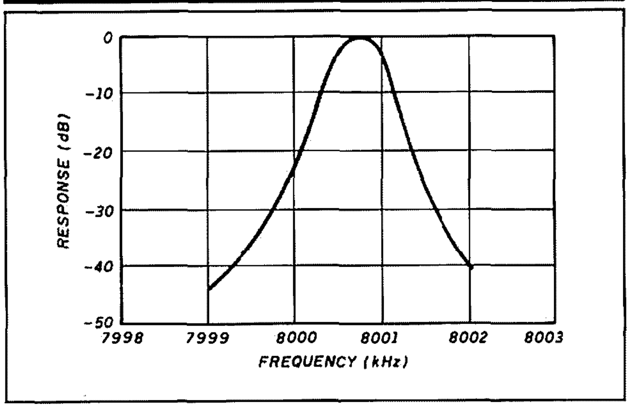
Measurement setup with trial filter.

FIGURE 8



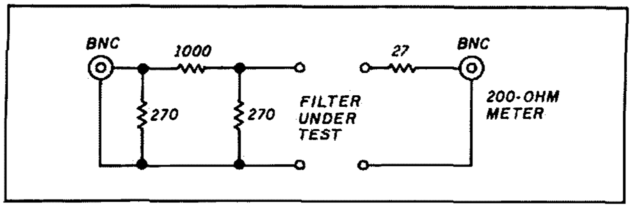
0.5-kHz CW filter.

FIGURE 9



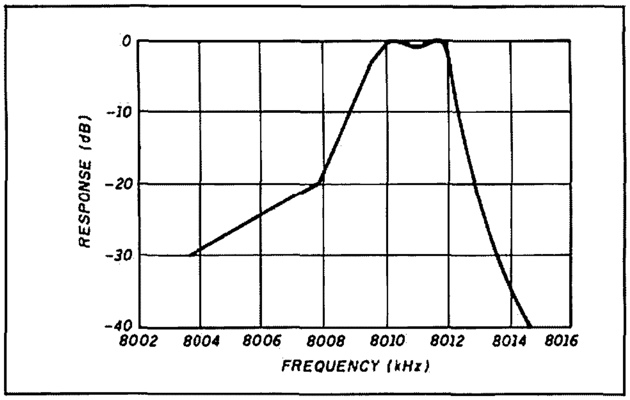
CW filter, calculated data.

FIGURE 10



220-ohm trial network.

FIGURE 11



SSB filter, measured data.

10 isn't possible with these low quality surplus crystals. There's barely a factor of 2 in this case. Nevertheless, with typical Amateur resourcefulness, I used the crystals anyway. As it turned out, they work just fine. However, the insertion loss is somewhat high.

The calculated resistance for the CW filter is stepped up to match that of the SSB filter with a shunt capacitor at the ends of the filter:

$$Cend = \left[\frac{1.59 \times 10^5}{Ro\ Fo}\right] \times \sqrt{\frac{Ro}{Rend} - 1} \quad -5\ (pF) \qquad (6)$$

where Ro is the desired (SSB) resistance. This equation derived by Hayward lets you keep the termination resistance constant and switch capacitor values only when going from SSB to CW bandwidths.

The filter shape and insertion loss aren't changed by this impedance step up. For my crystals, Cend = 267 pF. The filter circuit and its computed response are shown in Figures 8 and 9.

Trial measurements

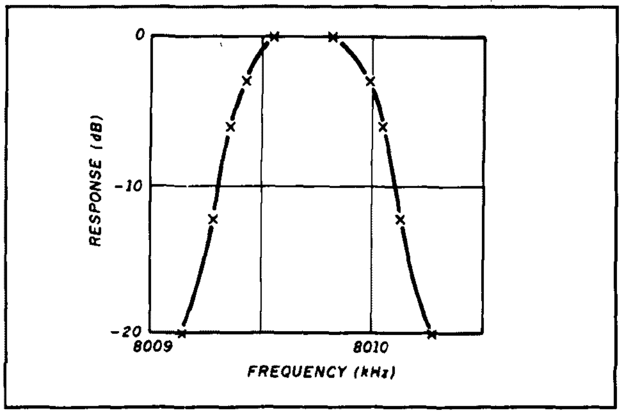
I verified the computed responses for both filters with the trial measurement setup shown in Photos D and E. I connected the crystals and capacitors for the SSB filter together on a piece of perforated circuit board by temporarily soldering the ends of their leads to the 220-ohm trial network (see Figure 10). This construction method leaves large inductive loops which could alter the filter characteristics. However, it didn't seem to have much effect at the 8-MHz frequency.

Take measurements by first adusting the signal generator for the peak response on the RF voltmeter. Add a fixed amount of attenuation with the step attenuator and note the new voltmeter reading. Then switch the attenuator back to 0 dB and adjust the signal generator above and below the peak frequency until the RF voltmeter reading is at the level you noted. Read these two frequencies on the frequency counter. Any error in the RF voltmeter calibration is immaterial with this procedure. Accuracy depends only on the attenuator accuracy and frequency counter calibration.

My step attenuator has 3, 6, 12, 20, 30, and 40-dB steps. The measured results for both the tuned and untuned (series capacitors shorted) SSB filter are shown in **Figure 11**. Correlation with the calculated data of **Figure 7** is very good.

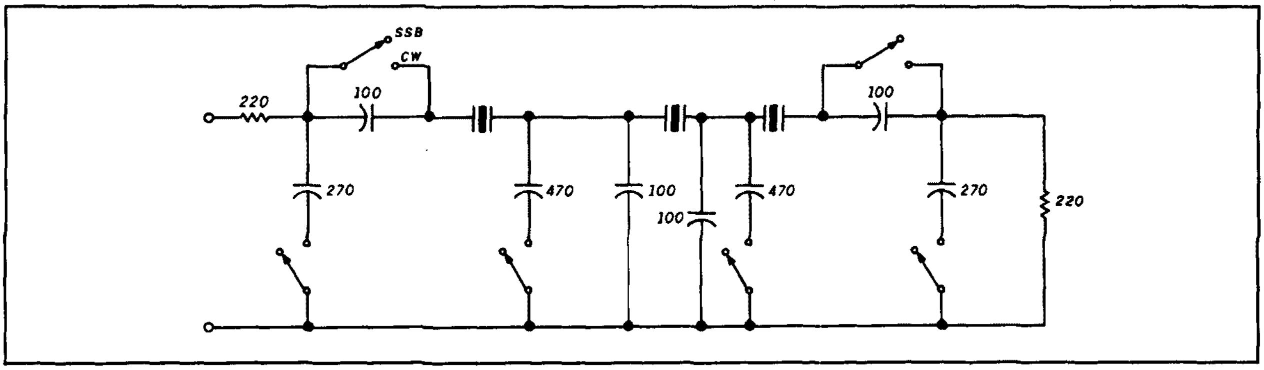
Now solder additional capacitors by their lead ends onto the SSB circuit to form the CW filter circuit. **Figure 12** shows the circuit's measured response. Again, there's good correlation with the calculated data (**Figure 9**). The measurements also confirm that low-Q crystals can be used to create narrow bandwidth filters.

FIGURE 12



CW filter, measured data.

FIGURE 13



Switched bandwidth circuit.

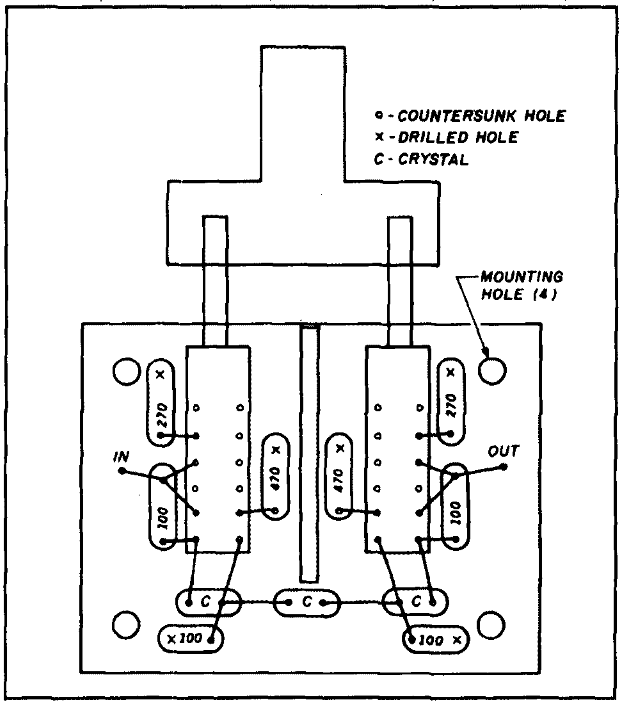
Constructing the switched filter

In a high frequency receiver, you must be able to switch from an SSB to a CW bandwidth. The overall switching circuit is shown in **Figure 13**. Six sets of switch contacts must close when going from SSB to CW. I used two four-pole push switches (GC Electronics catalog no. 35-492) and mounted them on a 2 × 2.5-inch piece of single-sided copper-clad circuit board. Two unused sets of contacts are available to operate indicator lights, adjust BFO frequency, or adjust the IF gain. As mentioned earlier, the insertion loss of the CW filter is about 8 dB greater than the SSB filter.

Component placement

Mount all components, including the switches, on the copper foil side of the board. Insert the leads in drilled holes, which are countersunk on the top side if the lead connection isn't connected to ground. This copper foil forms an excellent shielding ground plane. Insert component leads to ground into a drilled hole. Then solder the leads to the copper foil with a surrounding bead of solder on the top side. Interconnect the remaining component leads on the bottom side as required, to form the overall circuit.

FIGURE 14

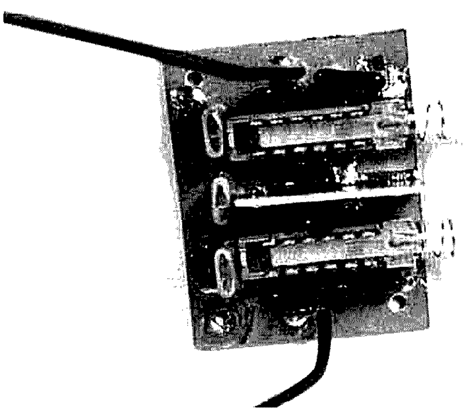


Component locations.

Component locations

Component locations are shown in Figure 14. Drill four large corner holes for mounting the filter behind the front panel in your receiver. Use short lengths of miniature coaxial cable for the input and output connections. Use a small piece cut from 1/4-inch Plexiglas™ to form a single pushbutton which lets you operate both switches simultaneously. Glue it to the push switches with plastic pipe cement. If you look at Photo F, you'll notice that I used two dipped silver mica capacitors where you might expect to see one. There's no magic here. I didn't have the correct value on hand and made it up with two other capacitors. There's no need to match capacitor values exactly; I didn't even attempt to do so. I simply soldered in standard value silver micas. Solder a shield between the two switches made from a 3/4 × 1-1/2 inch piece of the single board stock.

PHOTO F



Completed switchable bandwidth filter.

Final check

A final check of the switched filter duplicated the results measured in the trial construction. Even though the component lead length was considerably shorter, and there was shielding provided by the copper planes, I noted no differences.

Acknowledgments

I want to thank Wes Hayward, W7ZOI, for working out and publishing the design equations reproduced in this article. They made the job of constructing a crystal filter a straightforward procedure. Also I'd like to thank my associate Dr. Kischen Kapur, who doesn't have an Amateur license but likes to go to hamfests to buy and sell parts. Dr. Kapur supplied the crystals used in this article.

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2. J.A. Hardcastle, G3JIR, "Ladder Crystal Filter Design," *QST*, November 1980, pages 20-23
3. John Pivnichny, N2DCH, "A Homebrew Tuning Dial," *Ham Radio*, December 1988, pages 75-78
4. John Pivnichny, N2DCH, "A Sensitive RF Voltmeter," *Ham Radio*, July 1989, pages 62-67.
5. John Pivnichny, N2DCH, "High Impedance Rotary Step Attenuator," *Ham Radio*, February 1989, pages 24-26
6. Werner A. Schnider, "Verify Network Frequency Response with this Simple BASIC Program," *EDN*, October 5, 1977, pages 87-89.

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Ham Notebook

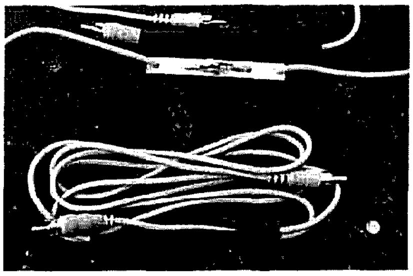
Simple Circuit Prevents Amplifier Control Relay Failure

The switching requirements of many linear amplifiers exceed the current rating of the relay in the driver stage. Immediate or premature relay failure can result when the contacts stick in the closed or transmit condition.

Once you've replaced the relay in the radio, your choices are buying an intermediate relay assembly (if you can find one), providing your own relay and wiring, or assembling a solid-state switch.

After looking into all the alternatives, I found my own solution. It's really quite simple. All you need is one transistor, an RCA terminated patch cable, a strip of metal, and some heatshrink tubing (see Photo A).

PHOTO A



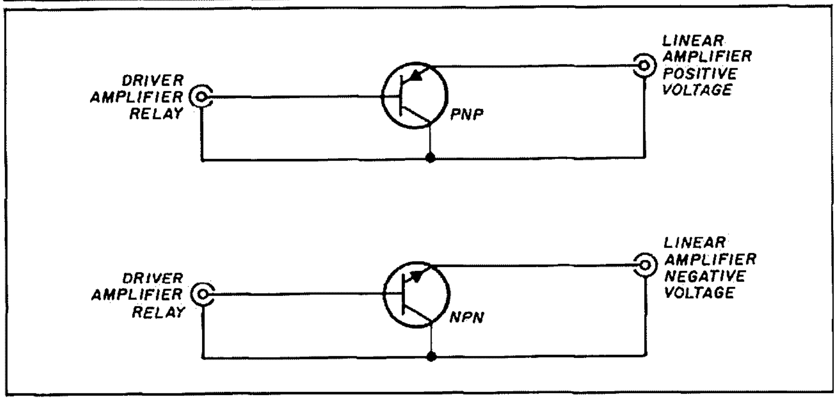
Simple hardware for providing protection to linear amplifier control relay.

The transistor

You'll need an NPN transistor for switching negative voltages or a PNP transistor for switching positive voltages, as required by your amplifier. The ideal transistor should have a low collector saturation voltage and be a high voltage switching type. Linear amplifiers seldom have transient suppression diodes on the relay coils. This causes spikes of many times the supplied voltage when the radio's relay opens. (Remember the voltage across a coil is related to the rate of change of the current running through it.)

I tried several different transistors and recommend MPSA44s or MPSA45s for switching negative voltages and 2N6519s or MPSA92s for switching positive voltages. The transistor receives all its current from the linear switch line and operates in saturation mode with less than 1 volt from emitter to collector, so very little power is dissipated. High voltage NPN transistors have higher gain than similar PNP transistors. This means the radio's relay is presented with one-fifth the switched current with PNP transistors and about one-twentieth the current with NPN transistors (which I used).

FIGURE 1



Solid-state switch protects relay in driver amplifier.

Construction

Drill two holes in each end of a 3/8 × 3-inch aluminum strip. Glue the transistor to the center of the metal strip with silicone seal. Cut the patch cable in the center and strip the wires for soldering. Stuff the stripped ends through the outer holes first and then through the inner ones to provide attachment to the metal strip. Solder the wires to the transistor for the circuit in Figure 1. If you were thinking ahead, you've already stuffed the heatshrink onto one of the cables so you can slide it over the assembly and shrink the tubing for protection. For added protection, cover the transistor and connections with silicone seal. I shrunk a small band on the linear's end plug, using blue for positive voltages (PNP) and black for negative voltages (NPN).

The switching cable is quiet and fast. It's simple to install and leaves your accessory socket free. You might consider using it to increase relay life, even where immediate or premature failure isn't evident.

The assembled NO7G+ or NO7G− cable is available for $15 plus $2 shipping and handling. Send your requests to me at 525 26th Avenue S., Seattle, Washington 98144.

David Smith, NO7G

Stop Your TH3 Junior Drooping

Recently I took possession of a TH3 Junior Yagi which was looking rather tired. I tried the old method of giving the tubing in the the elements half a turn and, although it looked better, it still drooped.

I made up three lubes (as in Figure 1) using the bolt in place of the anchor bolt in the element-to-boom bracket. The dowels are 5/8 inch in diameter and 12 inches long.

The braided rope (nonconductive) is 5/32 inch and is tied off outboard of the 15-meter traps (see Figure 2). Allow the rope to stretch under tension before putting the Yagi on the tower.

Perhaps this idea could be employed on the bigger Yagis if you use stronger materials. It works well on the TH3 Junior and deters large birds from perching on the elements.

By Arthur Brean, VK6SY

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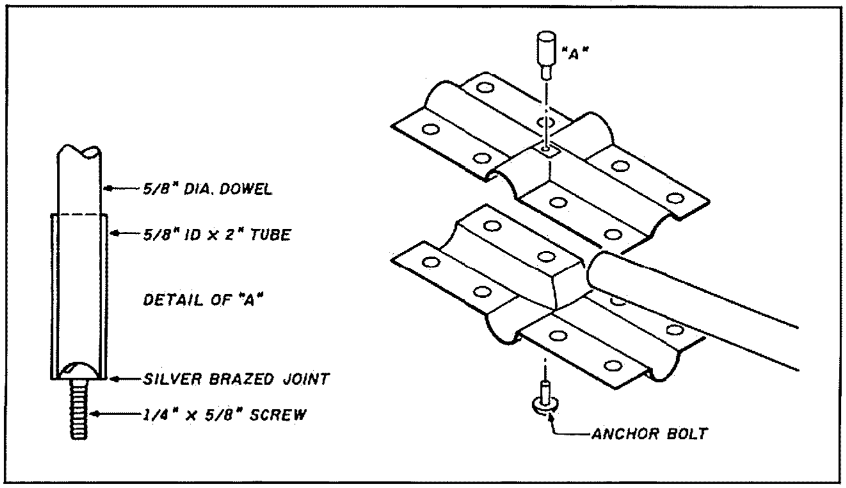
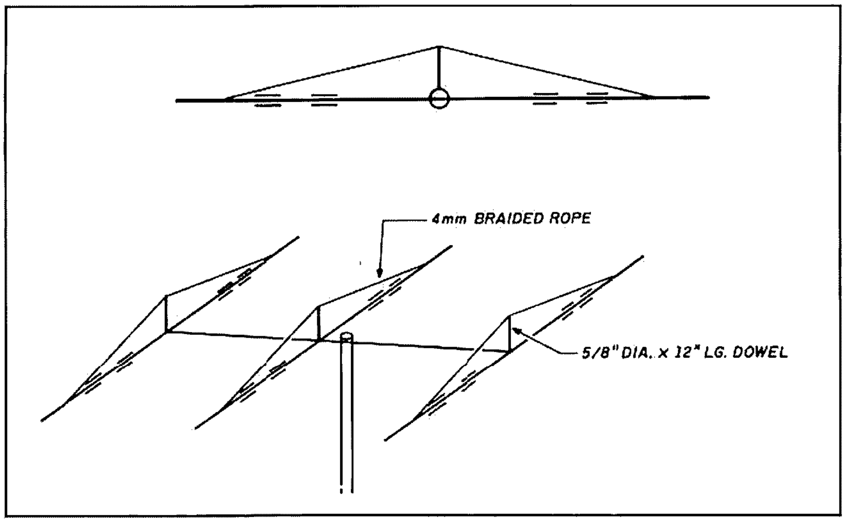
FIGURE 1



Boom-to-bracket element.

FIGURE 2



Braided rope (nonconductive) tied off outboard of the 15-meter traps.

INEXPENSIVE HALF-WAVE
2-METER MOBILE ANTENNA

By Glen Noble, WE7C, 4840 Schindler Road, Fallon, Nevada 89406

I needed a 2-meter mobile antenna, so I thought: "Why not build it myself? How hard could it be, anyway?" A search of the garage yielded an old gutter-mount CB antenna which could provide a mount and coax, but the whip looked hopeless for 2-meter use.

The hardware

I went to Radio Shack to look for a suitable whip and was pleasantly surprised when I found a stainless steel CB whip exactly the right size for a half-wave 2-meter antenna.

Now that I had the new half-wave whip and the old CB gutter-mount frame, my project was on a roll. While browsing at Radio Shack, I discovered that gutter-mount frames and other styles, like mirror mounts and single-hole types, were available at low cost without attached CB antennas.

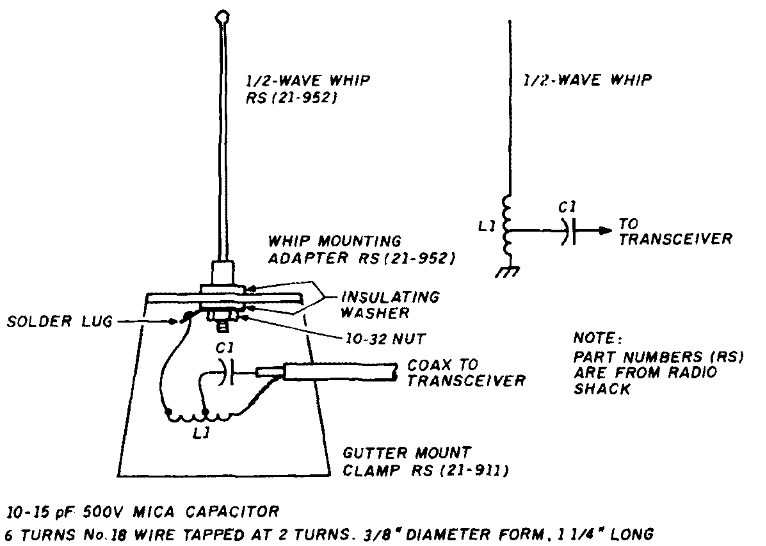
The impressive performance of my AEA "Hot Rod" half-wave handheld antenna heightened my enthusiasm for this project. I completed my design after consulting the RSGB *VHF/UHF Manual*.[1] It showed me how to turn my growing collection of parts into a usable antenna by adding a coil of wire and a small capacitor.

How it works

The main obstacle when building a half-wave whip antenna lies in matching its high input impedance to 50 ohms. Because the input impedance is high, you can't simply connect one end of the coax to the whip while grounding the other end, as you would for a quarter-wave antenna. However, if you add a six-turn coil tapped at two turns, along with a small capacitor, you can form a matching network that transforms this high impedance to the required 50 ohms.

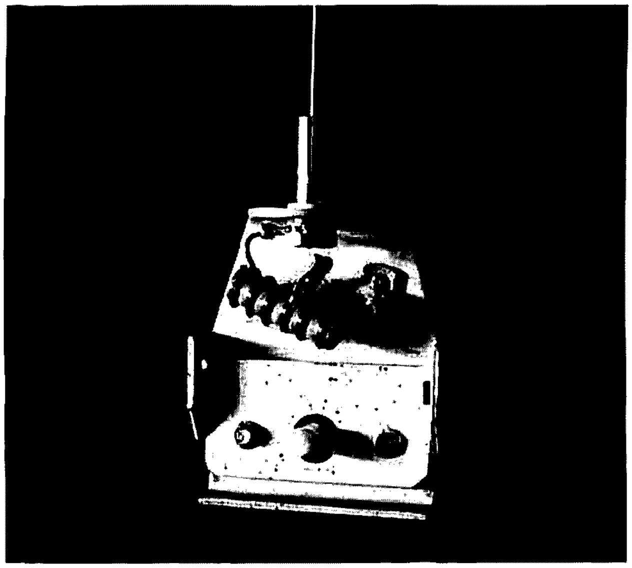
The coil is the principal means of transforming impedance; the capacitor compensates for a reactive component introduced by the coil. **Figure 1** shows the connections and parts used. The resulting antenna has good bandwidth and covers most of the 2-meter band with acceptable VSWR.

FIGURE 1



Pictorial and electrical schematic of the 2-meter half-wave antenna.

PHOTO A



Matching coil is located under the mount to inhibit radiation.

When building this antenna, I did deviate from the usual construction methods by locating the matching coil under the mount, where it can't radiate (see Photo A). This coil is usually placed above the mount, where it can contribute to signal strength on both transmit and receive. However, casual comparisons of this antenna with my AEA "Hot Rod," with the coil above the mount haven't revealed any major differences. Installing the coil under the mount does simplify construction significantly.

It's interesting that the half-wave antenna is thought to perform better than a 5/8-wave whip when a good ground plane isn't available. This could be quite helpful on a gutter-mount antenna, where half the ground plane is missing.

Construction

The stainless steel CB whip, Radio Shack part no. 21-952,* provides the 39-inches necessary for a half wave on the 2-meter band. My gutter mount was an old Radio Shack CB antenna similar to the current part no. 21-909. These parts cost about $6 each.

Start your construction by modifying or fabricating an insulating washer to insulate the base of the whip from the top of the mount. A similar washer is needed to insulate the nut and lug from the bottom of the mount. The top washer of my gutter mount had a lip that I fitted into the mounting hole to center the whip in the opening. The bottom washer was a typical flat plastic one. I had to thin these washers so the 10-32 threads on the whip would extend far enough through to allow me to attach a nut. I used a thin blade saw to cut one of the washers in half; you can also sand or grind them down. You may have to make these washers if you can't salvage them from a used antenna.

Next, wind the coil on a 3/8-inch diameter wooden dowel form. You need six turns of no. 18 wire, 1-1/4 inches long, with a tap at two turns. The series capacitor doesn't appear to be critical in terms of capacitance. The RSGB manual suggested 15 pF, but the 10 pF I had available worked fine.

*Replaces Radio Shack part nos. 21-904/908/940. Ed.

Finally, wire the assembly as shown in Figure 1.

Adjust the completed antenna for minimum VSWR. You can do this by compressing or expanding the coil turns, as well as adjusting the whip length. When you're satisfied with the VSWR over the band, use epoxy glue to cement the coil turns firmly in place and waterproof the wooden dowel form. It's worthwhile to monitor the VSWR when attaching the coil, as the epoxy may have a slight effect on the VSWR. Tweak the coil, if necessary, to optimize the VSWR before the epoxy hardens.

After the epoxy hardens, use a dab of silicone sealant to hold the whole coil in place and waterproof the coax. It's a good idea to leave a little extra whip length within the mounting adapter. This will allow for some adjustment when you mount the antenna on your the vehicle, or if the sealant has any effect on your VSWR. Photo A shows the completed antenna.

Closing remarks

With all the half-wave whips and mounts available, and all the used CB antennas lying around, it's easy to build VHF mobile antennas. It should also be possible to convert an AM/FM antenna using this approach. The Radio Shack half-wave whip screws into the base of the AM/FM antenna on my Toyota. The addition of a coil and capacitor at the base of the antenna in the fender well should produce good results on 2-meters.

The only real problems in converting CB antenna parts to VHF mobile use appear to come from adapting the various mounting thread sizes and the possibility of molded-in base loading coils.

Next time you need a VHF mobile antenna give this approach a try. In less than a day you can have the satisfaction of homebrewing a quality antenna using inexpensive, readily available parts.

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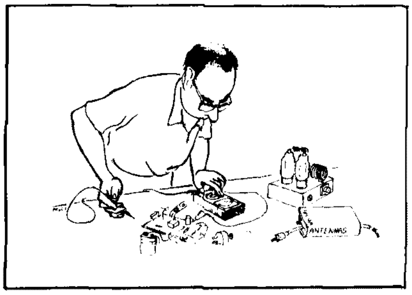
Ham Radio Techniques



By Bill Orr, W6SAI

THE MN ANALYSIS PROGRAM (THAT WAS THEN. THIS IS NOW)

In 1935 Marshall Mims, W5BDB, described a startling new beam antenna concept in *QST* magazine.[1] It was a two-element rotatable Yagi made of aluminum tubing elements. The antenna had remote control rotation and a direction-indicating system. Nothing like it had ever been seen before in Amateur Radio!

The Mims beam took over a year to perfect; four designs were built and discarded. Number five worked, after a fashion. Number six seemed better. By the seventh model, the beam was ready for action. It had quarter-wave spacing between the driven element and the reflector, and provided about 2.7-dBd gain and a front-to-back ratio of 10 dB.

Or so Mims thought. Running his 1935 design through a modern 1989 computer program shows that the pioneering ham band Yagi provided 4.3-dBd gain and a front-to-back ratio of about 12 dB. Mims made a pretty good beam and an excellent guess about its characteristics. His data was based upon measurements made by interested hams, using the S-meter of their receivers. (*Editor's note: Actually, "R-meters" were used in those days.*)

Antenna design and measurement techniques had a crude and shaky beginning but were refined over the decades. What was once thought to be rather routine field measurement proved quite tricky and complicated if reproducible antenna measurements were desired. Over the years the military and manufacturers built some large and expensive antenna ranges to measure the characteristics of HF and VHF antennas, but Amateur antenna measurements were rough-and-ready.

The computer enters the picture

The advent of the powerful digital computer soon provided a new insight into antenna operation. In 1968 an analysis technique known as "moment methods" was publicized.[2] This scheme dealt with the investigation of electromagnetic fields using computer techniques. Those who remember their high-school integral calculus will no doubt recognize the concept. But the new idea didn't catch fire until computer power was generally available at low cost.

The moment method provided the know-how to translate theory into practice. The job at hand was to provide a good computer analysis program for transmitting antennas.

The birth of the home computer brought a new level of antenna analysis. Some computerized methods for calculating antenna properties were based upon FORTRAN programs which used simple approximations for mutual and self-impedance to calculate element currents in arrays. The magnitude and phase of the element currents were then combined to produce moderately accurate radiation patterns. One of the most popular and well known of these programs was outlined and used by James Lawson, W2PV, in his series of articles[3] and his book.[4] Another contributor to this field was Stanley Jaffin, WB3BGU.[5]

About the same time, the Lawrence Livermore Laboratory at Livermore, California was manipulating a different, more powerful antenna analysis program — Numerical Electromagnetic Code (NEC) — a mainframe program which would be used to analyze electromagnetic fields.[6]

A derivation of the program (MININEC) was refined at the Naval Ocean Systems Center, Point Loma, California. It applied specifically to antenna analysis using IBM personal computers. Unfortunately, it wasn't user friendly and was too complex for everyday Amateur use. Even so, it was a gigantic step in the right direction.

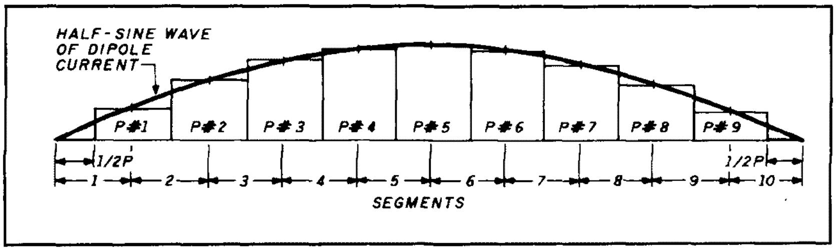
The K6STI MN antenna analysis program

Brian Beezley, K6STI, took the MININEC program and modified it for general Amateur work. He retained the original antenna modeling algorithm but optimized the code for higher performance. He then massaged the program to make it more applicable to Amateur service. The latest version of K6STI's program, MN 2.0, is the subject of this column.

The MN approach to antenna analysis can be illustrated by considering a straight conductor, like the dipole shown in **Figure 1**. The dipole is broken into segments for examination. The number of segments depends upon the complexity of the antenna and may be chosen by the user. In this case, ten segments are used. More segments would give higher accuracy to the computations, but would also increase computer time.

Each pair of wire segments defines a rectangular current pulse. As shown in the illustration, the current is modeled as uniform within each pulse. The current pulses are centered on segment boundaries and are the same length as the segments. The amplitude of the pulse closely approximates the amplitude of the current in the dipole at that point, as long as enough current pulses are used. The collection of pulses approximates the dipole current, which in this case is the classic half

FIGURE 1



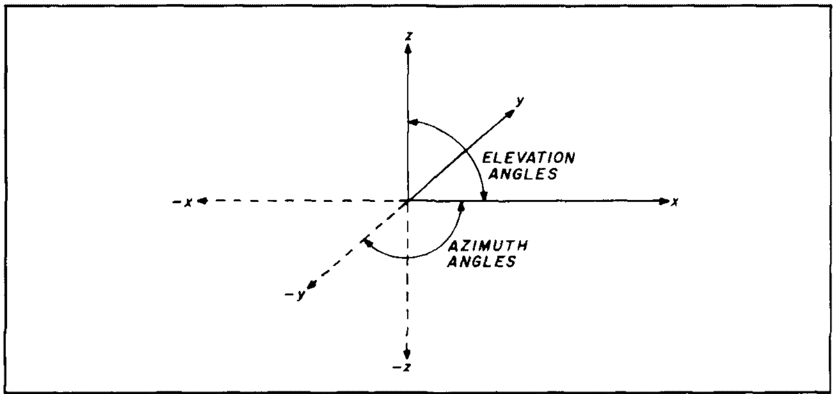
A plot of current in a dipole. The dipole is divided into ten segments, with nine full pulses centered on the segments. Half-pulses exist at ends of the dipole. The feedpoint is between segments 5 and 6 (pulse no. 5).

TABLE 1

Yagi Beam

20-Meter Yagi Antenna	
Boom	21.4 feet
Reflector	34.64 feet
Driven Element	33.2 feet
Director Element	31.34 feet
Element Diameter = 1 inch	

FIGURE 2



The X-Y-Z coordinate system. Three dimensions are shown. X-Y are vertical planes. Scale drawing of antenna may be placed anywhere in this frame of reference.

sine wave. A half pulse of zero current is placed at each end of the dipole; thus, there are ten segments and nine full pulses along the wire.

Figure 1 shows the picture of the wire that the MN program examines. However, certain restrictions apply. The wire is always straight. A bent wire is modeled by connecting two straight wires. Wires which cross or terminate at midpoints of other wires are modeled by defining separate wires which join. Connections are allowed only at wire end points.

The antenna feedpoint must be defined at a pulse. You can place the feedpoint where desired by specifying the correct number of segments in the wire. After doing it once or twice, you can almost make this placement by intuition.

Once all the data for an antenna is entered into an antenna file for the MN program, you can quickly determine antenna gain, front-to-back ratio, sidelobe level, beamwidth, feedpoint impedance, and vertical angle of radiation. These parameters once took hours of calculation or tedious field strength measurements to characterize.

As for the nitty-gritty of the program, let's run an exercise with MN on a three-element, 20-meter Yagi beam.

The antenna design program

You must define the antenna in terms that the MN program understands. Those who studied descriptive geometry in high school will grasp the idea immediately. Even if you didn't, you'll have no trouble picking up the procedure.

An X-Y-Z Cartesian coordinate system is used to refer to points in space (see Figure 2). The antenna is placed in this system. X and Y are the coordinates in the horizontal plane; Z is in the vertical plane. Think of the dimensions as X = length, Y = width, and Z = height. An antenna may be placed in the coordinate system in any position, but a directional antenna, like a Yagi, should be "aimed" in the +X direction. For convenience, the center point of the driven element is sometimes placed at the center of the coordinate system (X = 0, Y = 0, Z = 0).

The sample antenna is a Yagi of conventional dimensions as listed in Table 1. The Yagi is placed in the coordinate system shown in Figure 2. It's analyzed in free space so only the X-Y plane is required. The boom extends along the X axis and the elements fall along the Y axis (Figure 3).

Now that you've established the reference frame, you must describe the specific end points of the elements. The Yagi dimensions (in feet) are known, as are the end points of the elements. The midpoint of each element falls on the X = 0 axis and the end points fall in the +Y and −Y areas of the graph. For example, the reflector falls at a distance of X = −10.7 feet from the center point of the graph, the driven element is placed at X =− 1.28 feet from the center point, and the director is placed at X = +10.7 feet from the center point. This adds up to a 21.4-foot boom.

The reflector is 34.64 feet long; each half is 17.32 feet long. One end point of the reflector, which falls in the +Y quadrant, is labeled Y = +17.32 feet; the other end point, which falls in the −Y quadrant, is labeled Y = −17.32 feet.

The same sequence is followed by the end points of the driven element and the director. All the resulting end point plots are shown in the illustration.

All that remains is to tabulate the end points and choose the tubing diameter of the elements. One-inch tubing was chosen here for simplicity, although tapered telescoping elements may be modeled as well. The end points converted to our coordinate system are shown in Table 2. MN next needs to know the number and locations of feedpoints (sources), and the number of segments in each element.

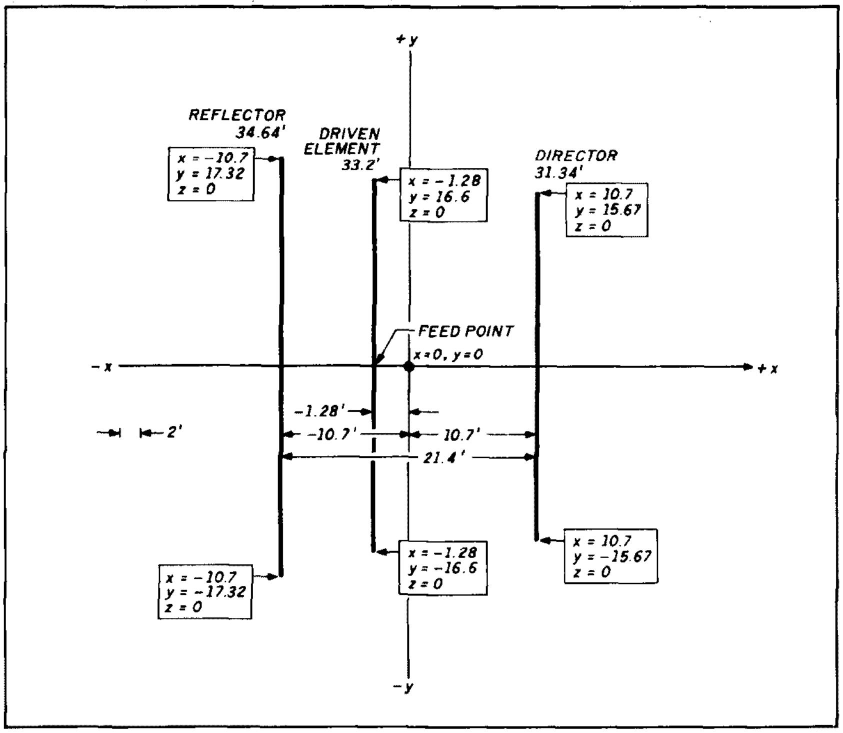
Ten segments are chosen for each element. This determines the number of pulses per element. The number of

TABLE 2

The X-Y coordinates of element ends.

Element	End point no. 1 Coordinates X	Y	Z	End point no. 2 Coordinates X	Y	Z
Reflector	-10.7	-17.32	0	-10.7	17.32	0
Driven	-1.28	-16.6	0	-1.28	16.6	0
Director	10.7	-15.67	0	10.7	15.67	0

FIGURE 3



The 20-meter beam laid out in X-Y coordinates.

TABLE 3

Antenna data file for input to MN. Line numbers are for reference only and shouldn't be in the final file.

1	3 element 20-meter Yagi			
2	Free space			
3	14.175 MHz			
4	Three wires, feet			
5	10	-10.7, -17.32, 0	-10.7, 17.32, 0	0.083
6	10	-1.28, -16.6, 0	-1.28, 16.6, 0	0.083
7	10	10.7, -15.67, 0	10.7, 15.67, 0	0.083
8	1 source			
9	14			
10	0 loads			

full pulses per element is the number of segments minus one. Thus, there are nine pulses in the reflector. There are also nine pulses each in the driven element and in the director. The source is located by counting the pulses, starting at one end of the first element of the antenna. In this case pulses are counted starting with the reflector, followed by the driven element. That is, nine reflector pulses plus one-half the number of driven element pulses specify the source. This falls at pulse 14 (nine reflector pulses plus five driven element pulses). Half pulses aren't counted. (If an odd number of pulses were chosen for the driven element, the feedpoint would be off center.)

Thus, the antenna has one source and the feedpoint is at pulse number 14. There are no loads in the antenna (inductors, capacitors, or resistors), so this portion of the MN file may be left out. It would be used if a triband Yagi with traps were being modeled.

Into the computer!

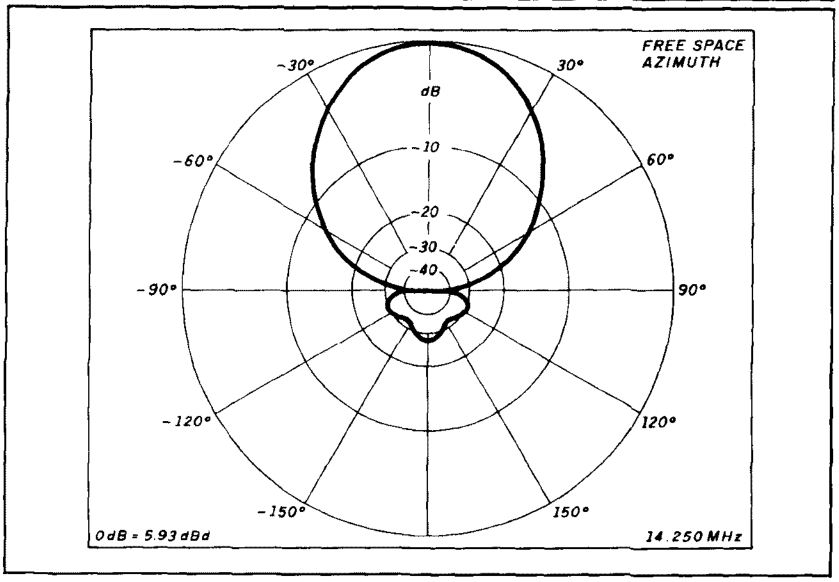
Are you with me? Now that we have all the information in hand, all that remains is to use an ASCII text editor or word processor to place it in a computer file in the stylized form shown in Table 3. Line numbers have been added for reference, but they shouldn't be in the final file.

Line 1 specifies a title for the antenna. This one is the "Three-element 20-Meter Yagi." Line 2 tells whether the antenna is in free space or modeled over the ground. I've chosen free space for this example, so MN looks for those words. If we had specified a Z dimension (like +45) to all elements, it would indicate that the antenna is 45 feet in the air.

Line 3 specifies the analysis frequency. In this case it will be 14.175 MHz, the center of the 20-meter band. Line 4 tells the number of "wires," or elements, and the unit of measurement. We're using three elements measured in feet. This completes the introductory phase.

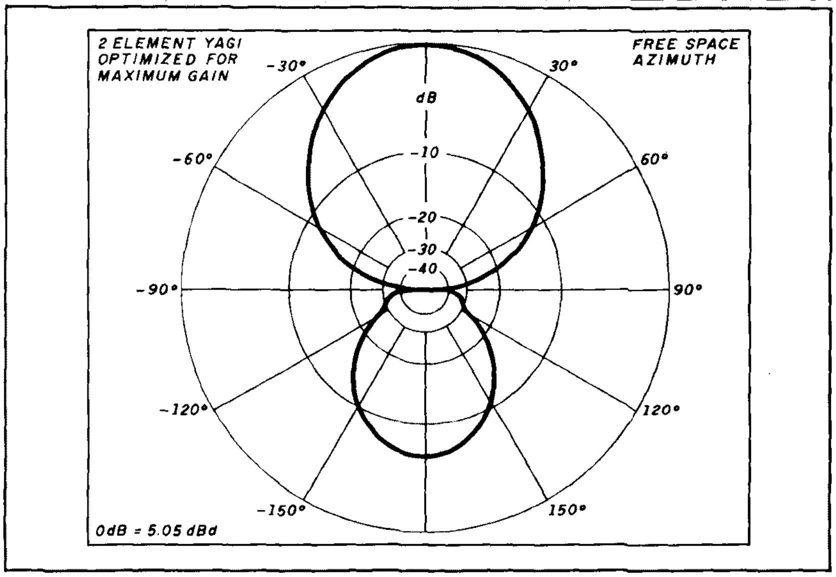
Now to the element specifics. Three computer lines are required (lines 4, 5, and 6), as there are three elements in the antenna. Each line (starting with the reflector) specifies the number of segments, the X, Y, Z coordinates of the wire (element) tips, and the wire diameter (chosen as 1 inch or 0.083 feet).

FIGURE 4



Azimuth plot of four-element Yagi by MN program.

FIGURE 5



MN plot of two-element Yagi shows good gain but poor front-to-back ratio. In comparing *Figures* 4 and 5 note that the gains for 0 dB (shown in the lower left) are different.

Line 5 lists the number of segments (10), the coordinates of the left and right-hand tips of the reflector, and the element diameter in feet. Lines 6 and 7 list the same for the driven element and director. Line 8 specifies the number of sources (1) and line 9 gives the drive pulse number (14). Line 9 can also set the applied voltage and current. But these values are optional and, if not entered, the computer picks a nominal default value to make things work out. Because there are no loads, line 10 is left out.

The MN program allows dimensions in feet, inches, millimeters, etc. and permits wire gauges to be used for element diameter. We needn't concern ourselves with these items now.

The completed antenna file is entered into the MN program, which is ready to run. I'll discuss the running of the MN program and show the results of this study in my next column. *Editor's note: There have been instances of (earlier versions) MN failing to load. This is characteristic of any big program which won't fit into available memory. When sufficient memory is provided, MN runs normally. The current version gives the message "out of memory" if it detects too little space.*

Antenna comparison with MN

MN can come up with some surprising answers. It's interesting to observe the power gain and field plot of a popular 20-meter four-element Yagi on a 26-foot boom. DXers favor this antenna. MN shows that the power gain of this array is 5.93 dBd with a front-to-back ratio of 27.17 dB at the design frequency. The field plot is shown in **Figure 4**.

A two-element 20-meter Yagi on an 8-foot boom is shown in **Figure 5** for comparison. The power gain at the design frequency is 5.05 dBd and the front-to-back ratio is 6.37 dB.

Consider that the boom of the four-element array is 18 feet longer than that of the little two-element beam. What did the extra 18 feet achieve? The big beam has a gain of only 0.88 dB over the small one. The real advantage is in front-to-back ratio — a whopping increase of 20.8 dB!

If you're only interested in power gain, it's a waste of space, time, and money to put up the big array. But if you need front-to-back ratio, there's no doubt as to which antenna is the logical choice.

After running various antenna designs on the MN program, I find it has proven decisively what I had known intuitively for many years — there is no free lunch. Bigger antennas are generally better in most respects than small ones. When you optimize a design for one characteristic (say, maximum gain), you suffer in another area (poorer front-to-back ratio or bandwidth). The "optimum" design depends upon your definition of optimum.

Fortunately, there's a computer program which can help you make some difficult optimization decisions. I'll discuss this interesting program "further down the log."

The "Dead Band" quiz — The winner is: K2KQU!

"A snowplow begins to clear a roadway at noon on a day during a steady snowfall. The plow moves two miles during the first hour and one mile during the second. What time did the snow begin to fall?"

This brain buster from Andy Loomis, KE0UL, really stirred up a mathematical storm! Andy found the problem in a C.R.E.I. electronics course text illustrating the use of differential equations. Runner-up Bob Phelps, NB7G, found the problem in *Differential Equations* by Agnew, 2nd edition, McGraw-Hill Book Company, New York. As Bob points out, the solution doesn't require the use of differential equations, but it does require calculus. Bob gets honorable mention for his concise explanation of the mathematics of snow plows and his insight into snowstorms.

Those snowplow drivers who realized that the speed of the plow at any instant is inversely proportional to the snow depth at that instant were on the right track!

Chief snowplow driver: Al Cohen, K2KQU

The snowfall began at 11:22:55 a.m., *or 0.618034 hours before noon.* Al notes that of all the quantities involved, only one (the snow depth) is a linear function. He says if the information given is placed in graph form, it's easy to jump to the conclusion that a graph of these points would place the starting time of the snowfall at 11:00 a.m., provided the plow moves 2 miles the first hour and 1 mile the second. Not so! The real graph of progress is a curve, not a linear function. Further, the speed of the plow is inversely proportional to the depth of snow, and the plow moves equal volumes of snow in equal periods of time.

Calling the speed **V**, the situation is expressed in the following form:

$$V = \frac{K}{C(t + H)} \qquad \textbf{(1)}$$

where **t** is the hours after noon and **H** is the hours before noon when the snowfall began. **C** is the rate of snowfall (a constant). **K** is a second constant which, hopefully, will disappear during the calculations.

Speed is related to distance traveled by integration over the time period, so **Equation 1** can be converted to an equation in "X" by integrating both sides:

$$\int_{t_1}^{t_2} V dt = \int_{t_1}^{t_2} \frac{K}{C(t + H)} dt \qquad \textbf{(2)}$$

$$= \frac{K}{C} \int_{t_1}^{t_2} \frac{dt}{t + H}$$

Because distance is the integral of speed, the integral of V from t_1 to t_2 is: $(X_2 - X_1)$.

On the right side, the integral of dt/(t + H) is the natural log of t + H. By simplifying:

$$(X_2 - X_1) = \frac{K}{C} Log_e \frac{(t_2 + H)}{(t_1 + H)} \qquad \textbf{(3)}$$

By further manipulation — substituting the given data for t_1 and t_2, and remembering that two times the log of a number equals the log of the square of the number — K drops out along the way, the logs drop out, and the result is:

$$\frac{1 + H}{H} = \left[\frac{(2 + H)}{(1 + H)} \right]^2 \qquad \textbf{(4)}$$

Expanding and clearing terms, the final formula is:

$$H^2 + H - 1 = 0 \qquad \textbf{(5)}$$

where H = number of hours before noon.

This can be solved by the quadratic formula to obtain:

$$H = \frac{-1 + \sqrt{5}}{2} \qquad \textbf{(6)}$$

which yields H = 0.618034 hours before noon, or 11 hours, 22 minutes, and 55.078 seconds.

Al points out that **Equation 5** and the value of H look familiar. The number H is one less than its reciprocal, which follows directly from dividing **Equation 5** by H:

$$H = \frac{1}{H} - 1 \qquad \textbf{(7)}$$

This is the "Golden Ratio" of Greek architecture — the ratio of length to width that's said to produce a rectangle most pleasing to the eye.

Finally, Al considers the "Fibonacci Series," a series of numbers that begins with 0,1. Each term thereafter is the sum of the two terms before it:

0,1,1,2,3,5,8,13,21,34,55,89......etc.

As this series gets longer, the ratio of two consecutive terms gets closer and closer to the "Golden Ratio" and to the term H in the snowplow problem. As interesting sidelights, the truncated series of only 12 terms has converged to 55/89 = 0.61798 and the 15th term of the series, 377, is the impedance of free space.

Al says this mysterious series of numbers shows up repeatedly in nature in such places as the pattern of markings on animals and the radial directions of bud sites on tree twigs! And, in some fashion, it also describes the physics of plows slogging through steady snowfalls!

Congratulations to Al Cohen, K2KQU, winner of the Dead Band quiz on snowplows!

Additional congratulations go to the following people who calculated the correct time: K0NT, K1KG, K6MV, NB7G, K0IXH, WD8KBW, AE2P, W4ZPS, W2IMU, KD9CM, WA2BSR/WB1GQU, KI6WX, K5RA, KB1FZ, NJ2P, KA1RCV, W6EL, G3RUH, W7HGS, and AH6HU.

Others who came close are: K5IU, KZ0Y, N4HUR, N4TXY, K7FC, NV7X, K9MFI, KA7D, W9WSS, KN6W, AI2S/NW2I, N0IBA, K4EF, N4TMI, WE8J, W7XU/0, W6TAD, N0ICS, ND6O, WE0F, W4UGW, W3EBY, NQ0V, K4KC, KA5MXX, K9AY, W1BG, KB0DON, KD8KF, K5ESV, KC8RF, N9DEO, W6EBY, K4ZLE, N2DT, KJ6GR, WA1SOO, WO8E, N6NS, NF7J, WB6JZY, K4WV, WB6LPS, KA8D, KI0F, KC6P, KB4LJV, KI7L, W7TLK, KA3ENQ, AA2V, W1BG, AA6CT, and W6JRZ.

Thanks all! I hope you enjoyed this little quiz.

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PERSONAL MESSAGE CENTER

Receive pages from your home station

By Roger Owens, AA4NX, 2042 Old Big Cove Road, Owens Cross Roads, Alabama 35763

My Personal Message Center (PMC) functions as a reverse autopatch, without the phone line, when connected to a radio transceiver.

Transmit page mode

I designed the PMC so I could receive pages from my home station. If another ham operator wants to reach me, he presses the "page" button on the PMC. This initiates a CW message (for example, AA4NX/RMT), which is generated every 10 seconds and stored in EPROM. If I don't respond within ten pages, the PMC terminates the call. If I'm available, I answer the page by entering a two-digit preselected access code. The code terminates the CW message, letting me activate the push-to-talk line to the transceiver. The microprocessor of the PMC does all the housekeeping functions. It doesn't allow PTT from the radio to be active for over 30 seconds at a time, and it identifies the station every 15 seconds of activity time on the transmitter. The call is terminated if the transmitter times out or the control operator enters a # sign.

Once the call termination (#) is received, the PMC identifies the station (AA4NX/SK) in CW, and disables the PTT to the transmitter. Two indicator LEDs operate during the page mode. One is the page LED, which blinks at a 1-second rate during the page mode. This tells you the device is active. The other LED, the PTT indicator, turns on whenever the PMC activates the radio transmitter.

Receive page mode

The PMC can also receive page commands from a mobile station. Enter your access code followed by the command *1. This activates the "callme" LED indicator which blinks at a 1-second rate, and a beep tone which is on for a second and off for a second. The PMC remains in this mode until it receives another *1 command or someone at the home station pushes the page button. Either of these commands terminates the callme LED and the beep tones. If the mode is terminated because the page button is activated, the PMC goes to transmit page mode operation.

DTMF squelch

The PMC acts as a DTMF controlled squelch. A set of relay contacts plugs into the external speaker audio jack, breaking the audio line. The DTMF squelch is activated any time a valid access code is received, and *during all* operating modes.

Remote site operation

If you need to make field measurements (after installing a new antenna system, for instance), you can access the PMC from your mobile or handheld using remote site operation mode. Activate this mode by entering your access code followed by *0. When the command is accepted, the PMC will:

- Activate the transmitter for 5 seconds of dead air time.
- Send the CW message AA4NX/RMT.
- Allow an additional 5 seconds of dead air time.
- Send the CW message AA4NX/SK.

After performing these tasks, the PMC returns automatically to its idle state.

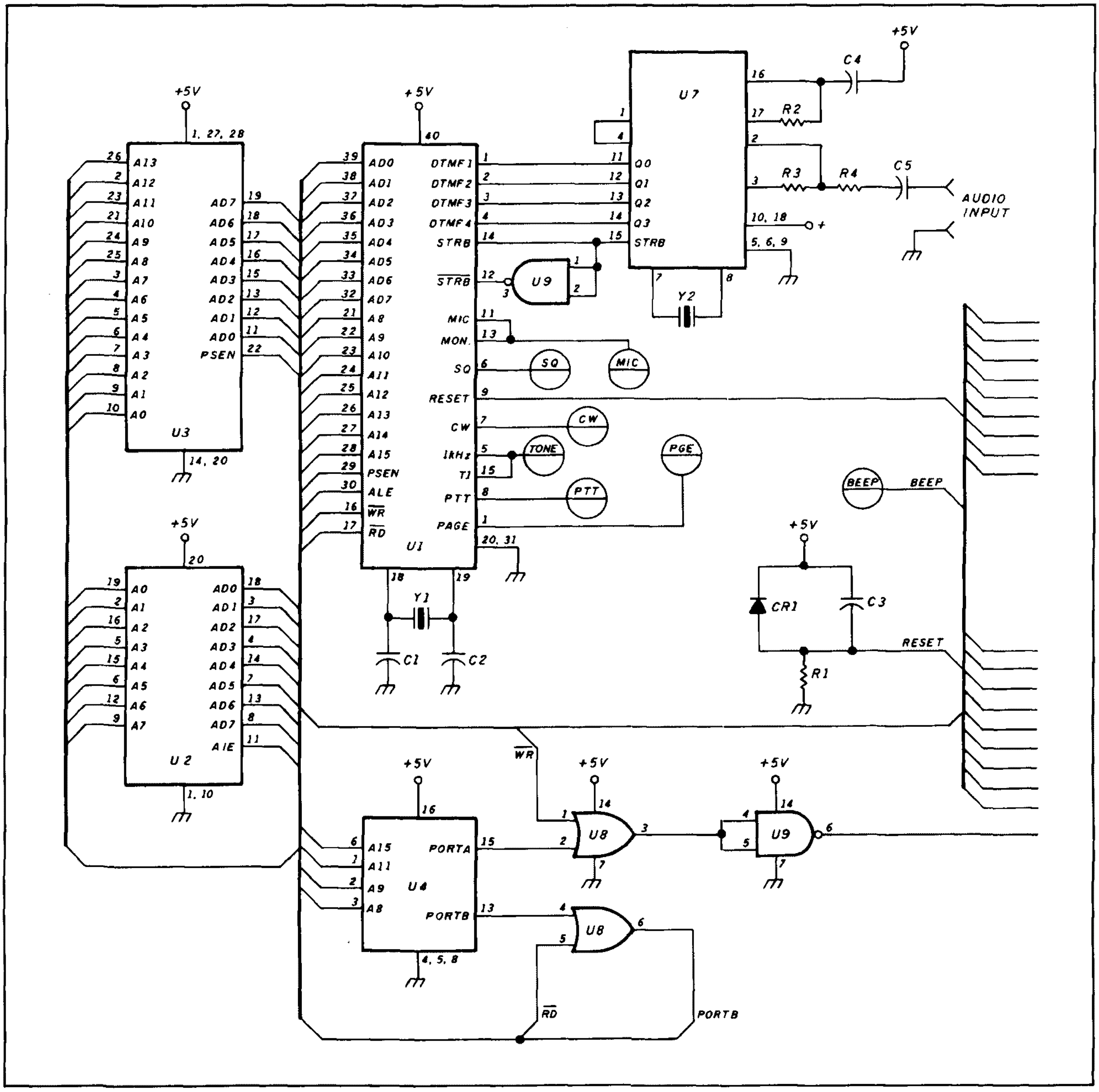
User activated ports

The PMC has four DTMF-activated outputs capable of controlling external devices. These outputs are for driving relays, etc., and are active high. To activate these ports enter your access code, followed by *2, *3, *4, or *5. To deactivate, repeat this step.

Radio interface

Bringing the receive audio out from your radio to the PMC requires just one modification. The best place to tap the audio is at the input to the volume control. At this point the audio

FIGURE 1A



Schematic of the Personal Message Center.

remains constant, regardless of the volume control setting.

Plug the DTMF squelch into the external speaker jack. Place a normally open switch (SW1) in parallel with the DTMF squelch relay contacts. To monitor the frequency, simply close the switch.

The transmitter audio line is unbroken from the microphone to the radio. Feed the Personal Message Center CW audio parallel with the microphone audio. This is a capacitively coupled output; it should not affect normal operation.

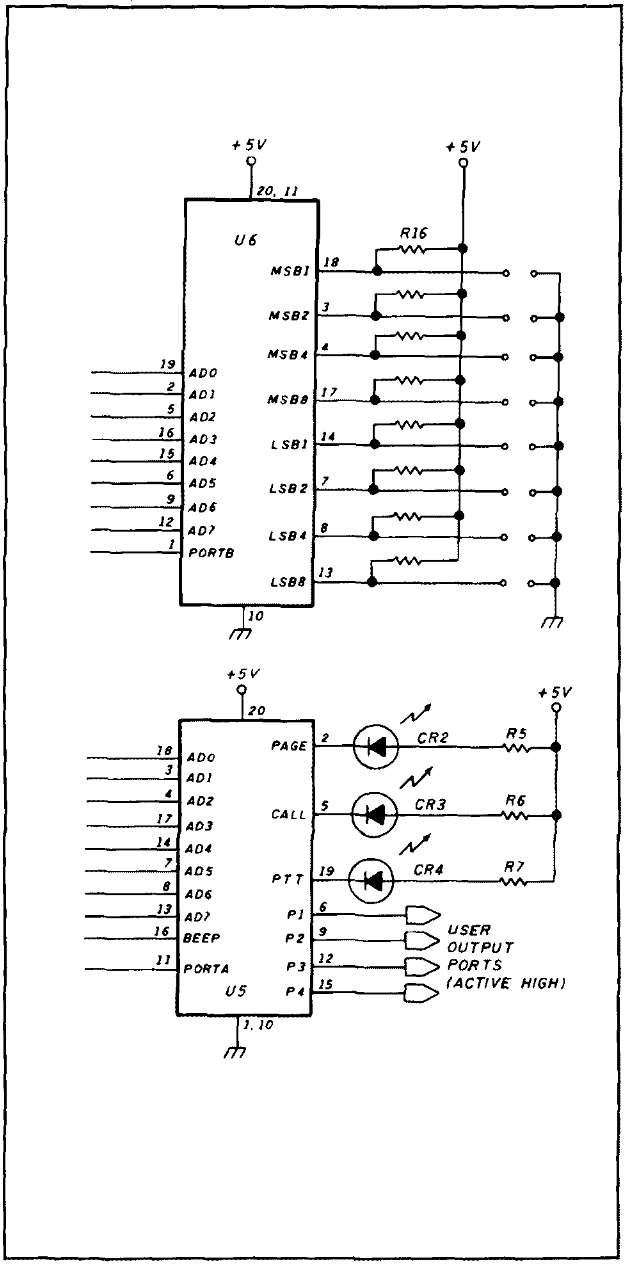
Break the push-to-talk line and feed it to a DPST switch (SW2). Attach one side of the switch to the PMC push-to-talk input. Connect the output relay contacts of the PMC to the push-to-talk input on the radio. This gives push-to-talk control to the remote station operator. Hook the other pole of the switch to the radio push-to-talk line for normal push-to-talk operation.

Circuit description

The Personal Message Center is basically a digital device (schematics shown in **Figures 1A** and **1B**). The only adjustment you'll need to make is to the beep level which connects to the microphone.

The microprocessor is an 80C31(U1) 12-MHz, 8-bit processor. The memory is a 27C64A 8K × 8 EPROM (U3) addressed by a 74HCT373 octal latch (U2). The system is reset on power-up by CR1, C3, and R1. External ports are accessed by the

FIGURE 1A



PARTS LIST

Designator	Part	Description	Quantity
CAPACITORS			
C1,C2		20 pF	2
C3		6.8μF 16 volt	1
C4,C5,C7,C9		0.1μF 50 volt	4
C6		220μF 16 volt	1
C8		1000μF 35 volt	1
INTEGRATED CIRCUITS			
U1	80C31	8-bit microprocessor	1
U2,U5,U6	74HCT373	8-bit latch	3
U3	27C64A	8K × 8 EPROM	1
U4	74HCT138	2 to 8 MUX	1
U7	8870	DTMF decoder	1
U8	74HCT32	Quad or gate	1
U9	74HCT00	Quad NAND gate	1
RESISTORS			
R1	8.2 k	1/8 watt, 1 percent resistor	1
R2	309 k	1/8 watt, 1 percent resistor	1
R3,R4,R14	100 k	1/8 watt, 1 percent resistor	3
R5,R6,R7	560 k	1/8 watt, 1 percent resistor	3
R8,R9	10 k	1/8 watt, 1 percent resistor	2
R11,R12	2 k	1/8 watt, 1 percent resistor	2
R13	100 k	1/8 watt, 1 percent resistor	1
R15	1 k	Pot	1
R16		10 SIP resistor	1
SEMICONDUCTORS			
CR1,CR2, CR3,CR4, CR6,CR7	1N4148	G.P diode	6
CR5	DF04	Bridge rectifier	1
Q1,Q2	2N3904	NPN transistor	2
Q3	2N3906	PNP transistor	1
Q4	7805	+5 volt regulator	1
RELAY (Miscellaneous)			
RLY 1,2		1A05	2

decoder 74HCT138 (U4) and associated logic gates. The two-digit access code is read through a 74HCT373 (U6). The actual selection is made by shorting or opening the appropriate bit. Each is binary weighted. The first four positions select the most significant digit of the access code; the second digit is selected by the upper four.

Example:

Shorting position	1 2 3 4	5 6 7 8
Open/short	O O O S	O O S S
Binary	1 1 1 0	1 1 0 0
Access code	7	3

The output control port 74HCT373 (U5) handles the page LED, beep, and user control functions.

The beep output control pin and a NAND gate (U14) drive PNP transistor Q1, which produces the drive for an 8-ohm 2-inch speaker. The 1-kHz tone signal is generated by the microprocessor. This tone is also applied to another NAND gate (U14), activated by the microprocessor CW control bit. The output of this gate supplies CW audio to the transmitter through its associated components.

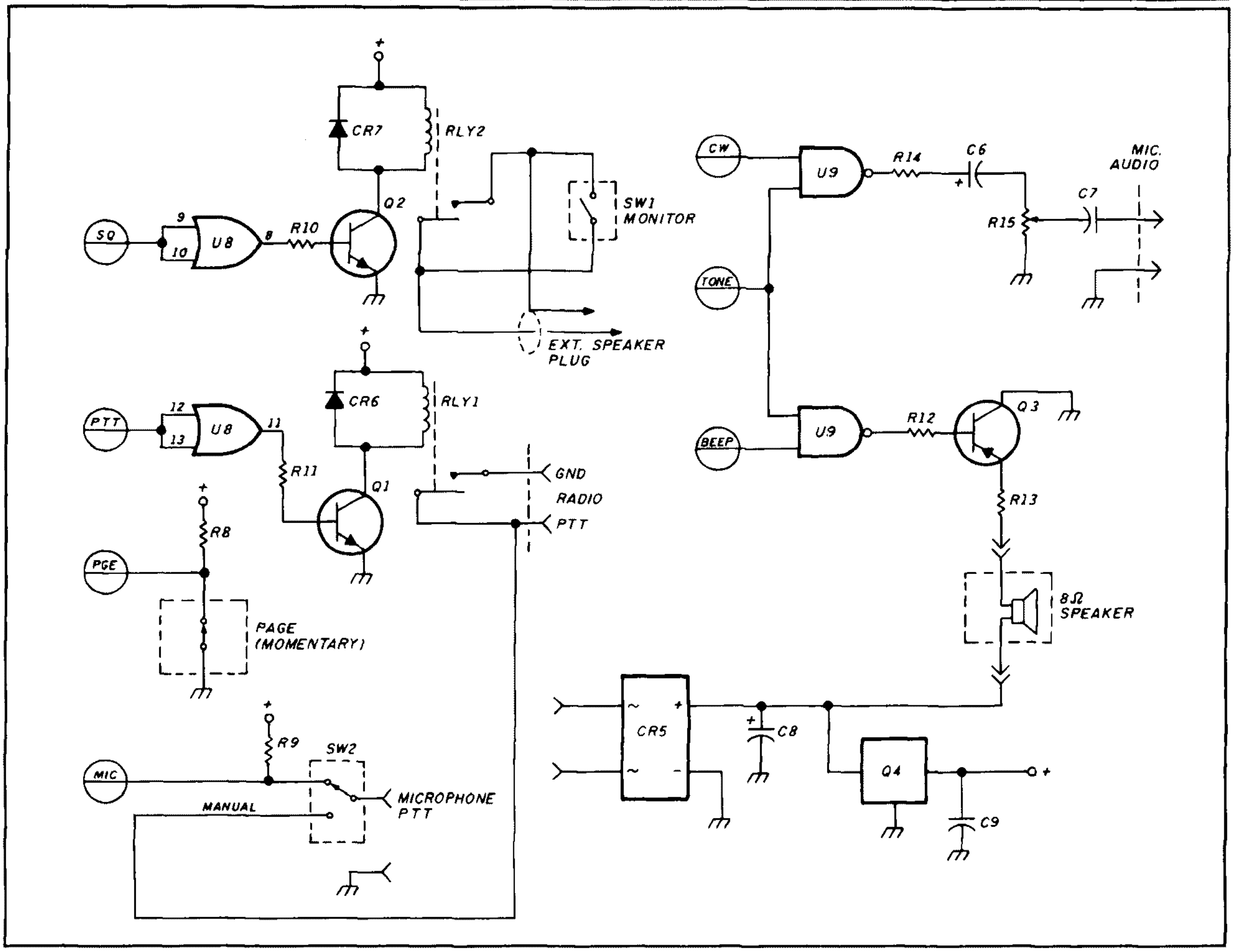
The DTMF receiver is a GTE 8870 device (U7). Receive audio is applied to the chip through C5, R3, and R4. Output detection delay is controlled by C4 and R2. When valid DTMF tone is detected, a strobe is applied to the microprocessor, along with a 4-bit code. The microprocessor decodes this information and performs the task requested.

The microprocessor controls the activation of both the push-to-talk and DTMF control squelch relays. Both circuits are identical and are buffered by OR gates (U8).

The push-to-talk line supplied from the microphone switch, which is active low, is a microprocessor input control line. The page push-button switch which uses a normally closed (momentary) push button is another. This furnishes an active high signal to the microprocessor to initiate the paging sequence.

The power supply consists of a rectifier bridge (CR5), filter capacitor (C8, C9), and +5 volt regulator (Q4). Complete, assembled pc boards are available from Valley Communica-

FIGURE 1B



Additional detail of the schematic of the Personal Message center.

tions, P.O. Box 277, Owens Cross Roads, Alabama 35763. (Write for price information.) PC boards with EPROM cost $27. Programmed EPROMS are $8.00. (Please specify call.) Other components may be purchased separately.

Note: Carefully review FCC Rules concerning remote control operation (Part 97.79) and reverse autopatches. This device may not be operated below 220.1 MHz under FCC rules concerning remote control links. *Ed.*

A DEEP NOTCH RESONANT FILTER

Filter using a 2N2222 transistor provides notch greater than 30 dB

By Douglas A. Kohl, WØTHM, 417 6th Avenue, N.E., Osseo, Minnesota 55369

The ideal notch filter passes zero signal voltage over a very narrow bandwidth. The signal at all other frequencies will pass through the filter unaffected. In practice, there will always be some signal voltage that passes through the filter at the notch frequencies.

One of the techniques commonly employed in active notch filters using op amps subtracts a portion of the input signal from the output of a frequency modified stage.[1] When a fraction of the input signal is matched to equal the voltage present at the frequency of the notch, the depth of the overall filter output at the notch frequency will increase and the output may approach zero voltage.

The same principle can be applied to resonant LC filters, which can work at radio frequencies much higher than most op amp notch filters. The simple series-resonant filter shown in **Figure 1A** exhibits a low output voltage, e_o, at the resonant frequency:

$$f_{notch} = \frac{1}{2\pi \sqrt{L\,C}}$$

At resonance the notch voltage is *not zero*, due to the effective wire resistance (R_w) in the inductor and other coil/capacitor losses. The value e_o results from the voltage divider action of R_1 and R_w (see **Figure 1B**), where e_g is the input signal voltage applied to the filter.

Note that the capacitor and inductor reactances cancel at resonance, leaving only R_w effectively in the circuit. Below the resonant frequency, the capacitive reactance becomes much larger than R_w. The divider fraction in the formula approaches unity, making e_o nearly equal to the input. Similarly, at frequencies above resonance, the inductive reactance increases and e_o increases (see **Figure 1C**).

Q versus notch depth

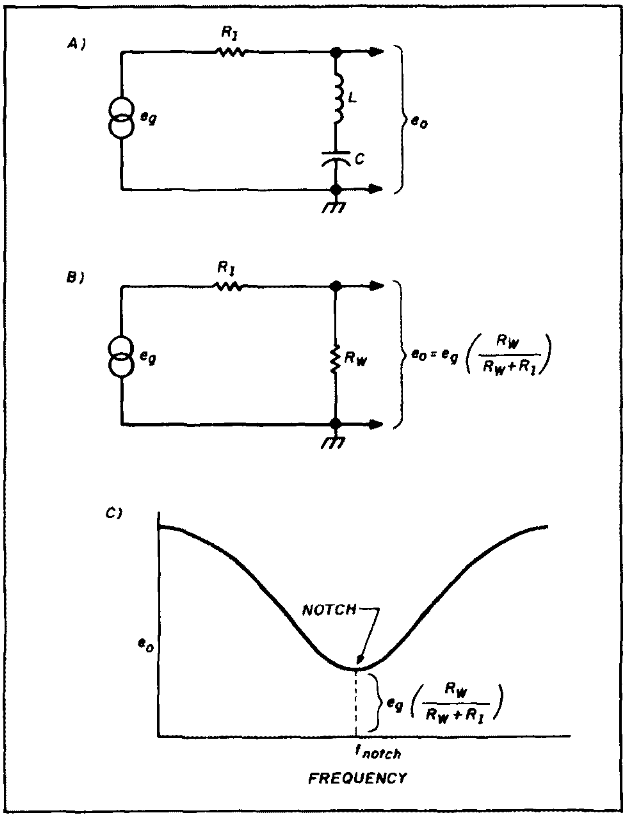
In a series-resonant circuit, the current that flows at resonance must pass through coupling resistor R_1 as well as the source of voltage, which has some internal resistance (see **Figure 1A**).These additional resistances *lower* the Q of the circuit accordingly:

$$Q = \frac{\textit{inductive reactance}}{\textit{all circuit losses}} = \frac{2\pi\, F_{notch}\, L}{R_w + R_1 + R_g} \quad \textbf{(1)}$$

where: R_g is the internal resistance of the signal source.

To increase the circuit Q and make the notch narrower in frequency,* an additional resistor (R_2) is added to provide a low resistance path for the circulating current between the capacitor and inductor at resonance. In Fig-

FIGURE 1



A conventional series-resonant filter and equivalent circuits.

$$^{*}\; Q = \frac{F_{notch}}{bandwidth}$$

ure 2A, the smaller R_2 is made the higher the Q. At resonance, the circuit acts as if coil resistance R_w is in parallel with R_2 (**Figure 2B**), which can be combined into one equivalent resistance, R′, for analysis (**Figure 2C**).

To make the filter input resistance high, you'd choose R_1 to be much larger than R_2. Thus:

$$e_o = e_g \frac{R'}{R_1} \qquad \textbf{(2)}$$

This would make e_o in the notch much smaller than e_g.

The results of the circuit in **Figure 2A** are shown in **Figure 3**. The bandwidth of the notch is the smallest for the highest Q condition (R_2 = 10 ohms). However, the output voltage at resonance is relatively high; that is, the notch depth is shallow.

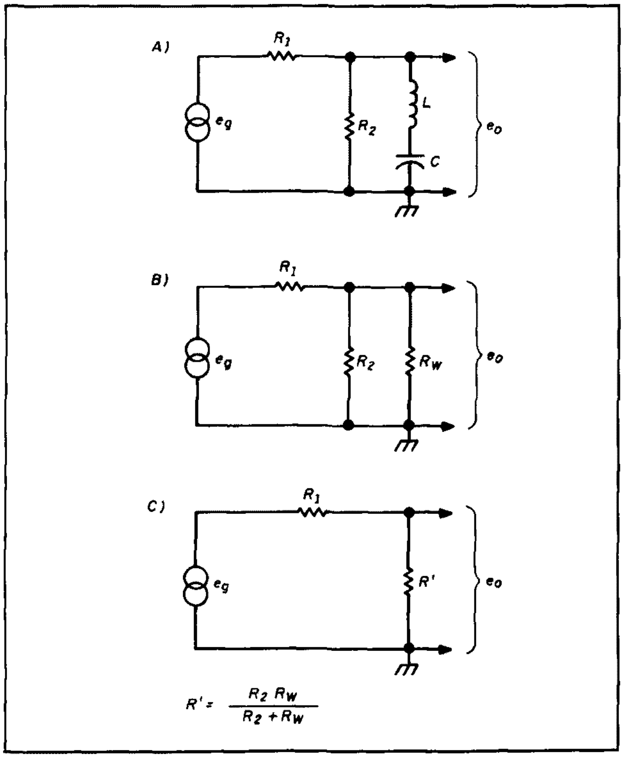
When R_2 = 100 ohms, the circuit Q decreases to 9, but the output voltage at the notch is much smaller. The circuit design problem becomes either a narrow bandwidth with a shallow notch, or a wide bandwidth with a deep notch. The circuit shown in **Figure 4** solves the problem and has both a narrow bandwidth and a deep notch.

The circuit

Figure 4 contains the series-resonant circuit discussed previously with e_o, the voltage drop across it, identified. When R_2 is small to achieve high Q, e_o at resonance is about 85 percent at frequencies above and below the notch. See the upper curve of the normalized frequency response shown in **Figure 5**. Even though it has very high Q, the depth of the notch is so shallow that it would be a very poor filter without the rest of the circuit. The phase of the notch voltage at e_o is the same as that of the filter input signal.

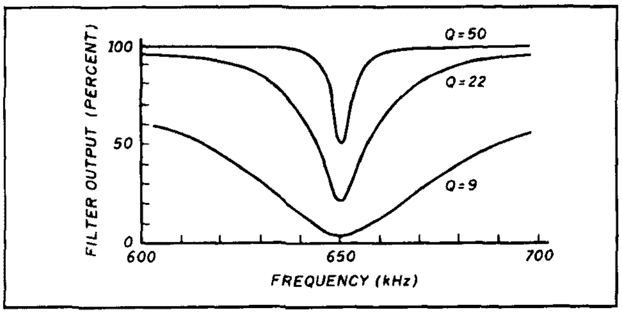
A 2N2222 transistor extracts the difference between the input signal and e_o, canceling e_o from the output at the

FIGURE 2



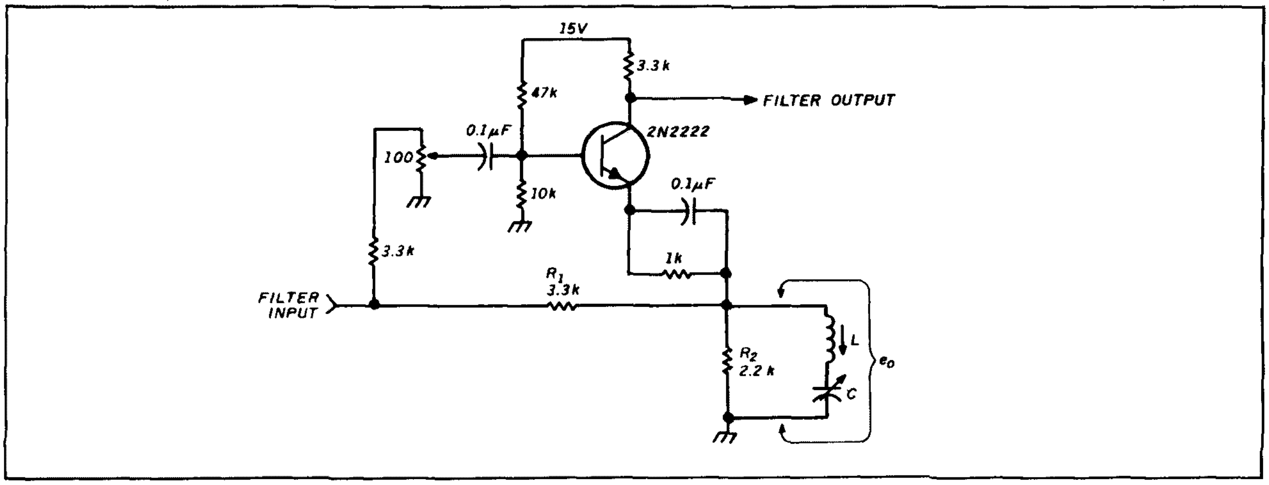
High Q series-resonant filter and equivalent circuits at resonance.

FIGURE 3



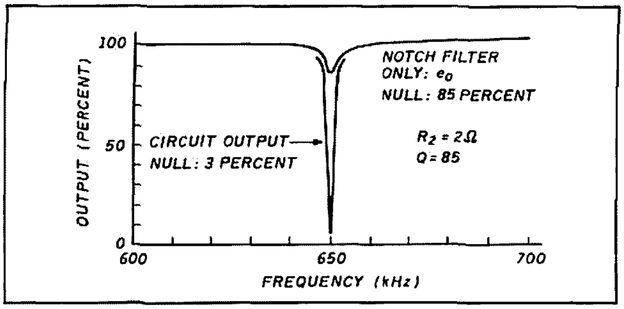
Frequency response of high Q series-resonant filters.

FIGURE 4



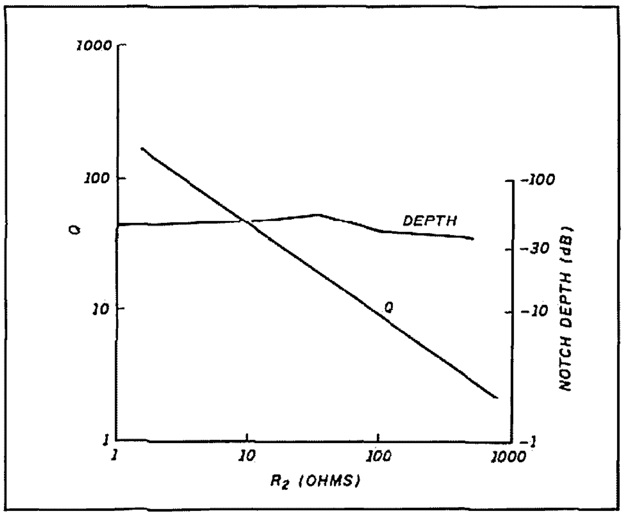
Schematic diagram of notch filter.

FIGURE 5



Depth of notch improvement of the circuit.

FIGURE 6



Circuit performance for Q values from 8 to 98.

notch frequency. To do this, use a pot to adjust the amount of the input signal appearing at the base of the transistor to match exactly the e_o voltage present at the emitter at resonance. The filter output voltage at the notch center will typically drop to less than 3 percent. See the lower curve of Figure 5 for particulars.

To change the notch frequency, you may change either L or C. Because the Q varies with frequency, you'll also need to readjust the pot to achieve the deepest null performance for each different frequency.

Results

The maximum Q attainable depends on the loss characteristics of both the inductor and capacitor you chose. Experimental data presented in Figure 6 show the effect on the circuit Q as R_2 is varied. The coil is an air-core RF inductor.

You can adjust the circuit to provide a deep notch exceeding 30 dB over a wide range of Q values or bandwidths. The voltage divider loss at the series resonant circuit, e_o, was restored at the filter circuit output as a result of transistor gain.

REFERENCES

1. Frank R. Dungan, *Linear Integrated Circuits for Technicians*, Breton Publishers, North Scituate, Massachusetts, 1984, page 192-3.

EFFECTIVE NOISE TEMPERATURE

PART 1: INTRODUCTION AND BACKGROUND

Understanding optimum noise performance

By Michael E. Gruchalla, P.E., 4816 Palo Duro NE, Albuquerque, New Mexico 87110

The concept of effective noise temperature is used to specify noise characteristics of low noise amplifiers and other low noise systems. It's generally understood that the lower the value of the effective noise temperature, the "better" the amplifier. Just exactly what is the effective noise temperature and why is it used instead of the more traditional noise figure specification? I'd like to discuss how the effective noise temperature specification is defined, its relation to the noise figure, and some of the pitfalls that you may encounter when trying to apply this specification.

Thermal noise

First, you must look at thermal noise and how it affects an electronic circuit. In 1928, Johnson showed that noise power is available in a system simply by virtue of the system temperature.[1] This is thermal noise, often called Johnson Noise because of his initial investigation of the phenomenon. Johnson showed that the noise power available to a *matched load* from a thermal source is a function of the absolute temperature, the noise bandwidth, and Boltzmann's constant (**Equation 1**).

$$P_n = k\ T\ BW_n \qquad \textbf{(1)}$$

where:

P_n = noise power (W) available to the load
k = Boltzmann's constant
= 1.38×10^{-23} W-s/°K
T = Absolute temperature (°K)
BW_n = Noise bandwidth (Hz)
= $\pi/2$ × signal bandwidth (see text)

Only resistances contribute this thermal noise; ideal reactive components don't. However, all real devices exhibit some resistive component which contributes thermal noise power to a circuit.

Spot noise

Often the noise is specified in terms of a 1-Hz noise bandwidth — noise power per Hz, noise voltage, or current per root-Hz. This is called the spot noise and is given by **Equation 2**, which is simply **Equation 1** with a 1-Hz noise bandwidth.

$$P_n(spot) = k\ T \qquad \textbf{(2)}$$

It's important to remember that the noise performance of many devices — like bipolar transistors, FETs, and amplifier diodes — is often specified *in* terms of spot noise at various operating frequencies. This information is provided because the noise performance of typical devices isn't the same at all operating frequencies. If you use a spot noise value alone for noise computations, without the noise bandwidth of the application, you'll get a noise level much lower than the level that can actually be obtained.

Uncorrelated noise voltages

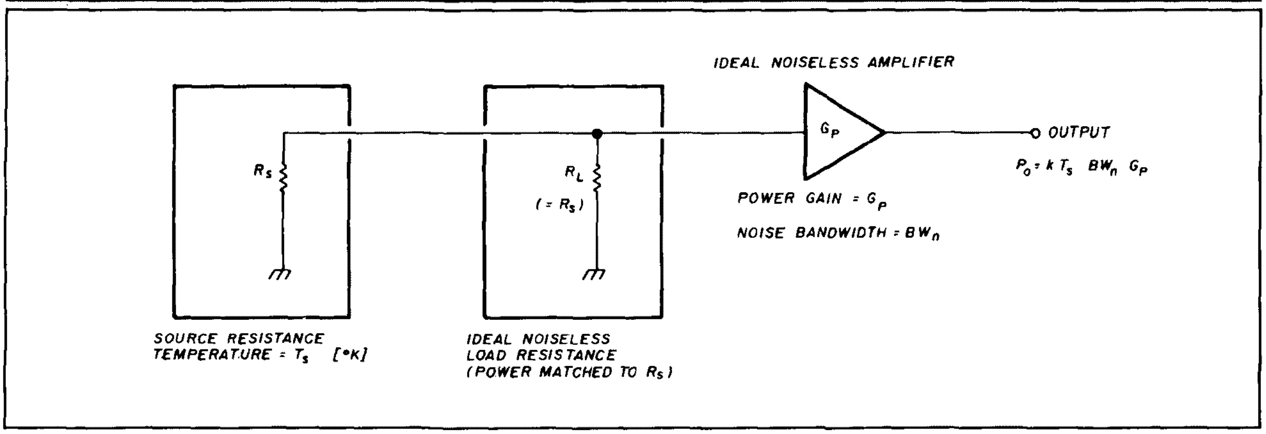
Noise is a statistical quantity. As such, noise voltages and currents can't be combined by simple addition. Only the mean square values may be added. The mean square value is simply the square of the rms value. Because noise powers are functions of the mean square of noise voltages or currents (for example, power is equal to the rms voltage squared divided by a resistance), noise powers may be added directly where necessary. However, when unrelated (*uncorrelated*) noise voltages (or currents) are to be added, the combined voltage is the square root of the sum of the mean squares of the individual voltages. For instance, if you have a 1-mV rms noise source in series with a 2-mV rms noise source, the total noise voltage due to the two sources in series is 2.24 mV rms (the square root of 5 mV),* not 3 mV.

Noise bandwidth

Noise bandwidth is the bandwidth used in noise computations. It's shown here as BW_n. The noise bandwidth is related directly to the signal bandwidth and the nature of the transfer function of the system under consideration. (The mathematics involved in the development of the noise bandwidth is beyond my intent here. For those interested

* $\sqrt{5\ mV_2} = 5 \quad \sqrt{5\ mV} = 2.24$ Ed.

FIGURE 1



"Warm" resistive source with matched load.

in digging deeper, **Reference 1** is a very good introductory text on noise phenomena.) When the amplifier has a relatively flat passband response, and the bandpass cutoff characteristic follows a single time constant characteristic or rolls off at 20 dB/decade, the noise bandwidth is very nearly $\pi/2$ times the signal bandwidth. This approximation is generally sufficient for many noise calculations. In cases of very low noise systems and very accurate noise calculations, the exact noise bandwidth must be measured for the specific system to obtain accurate computations. For this article, I used a single time constant cutoff and the noise bandwidth is defined as $\pi/2 \times$ the signal bandwidth.

Temperature definitions

The temperature in **Equation 1** is the absolute temperature in degrees Kelvin. The absolute temperature in °K is 273° greater than the temperature in °C. Absolute zero is 0°K and −273°C. The value of 290°K (17°C) is commonly used as nominal room temperature for electronic applications.

Thermal noise power is the result of atomic thermal agitation of a system. This is a basic principle of physics and can't be altered by circuit design or other techniques. The only way to reduce the thermal noise power in a specified bandwidth is to reduce the temperature.

The "Ultraviolet Catastrophe"

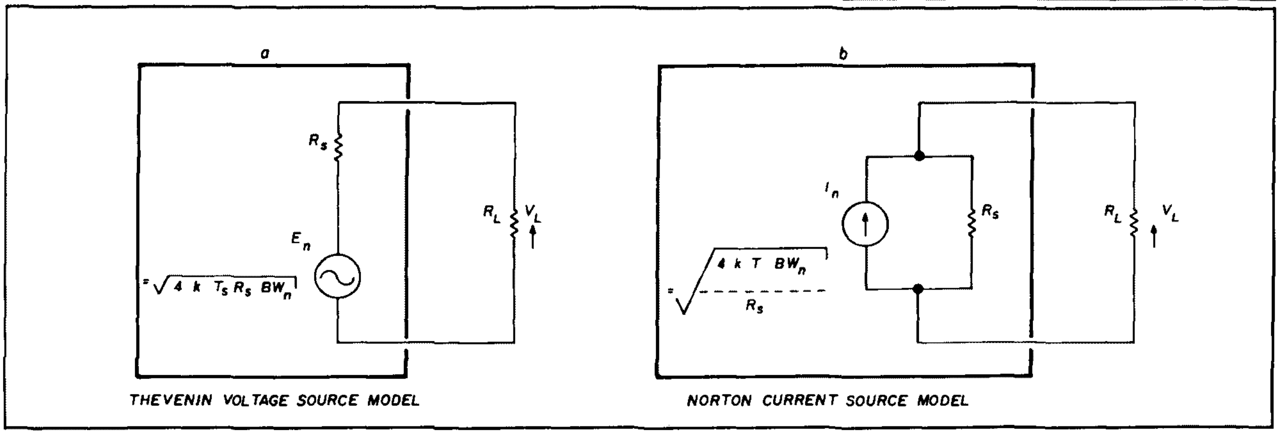
Equation 1 implies an interesting paradox. If you make the bandwidth infinitely wide, **Equation 1** suggests that the available noise power will also be infinite. In early noise research, this led to what was termed the "Ultraviolet Catastrophe."[2] This paradox came about because the classic mechanics of the time couldn't explain the noise processes at very high frequencies (ultraviolet wavelengths). The advent of quantum mechanical understanding led to principles which showed that the available noise power per Hz is reduced at very high frequencies. However, **Equation 1** is quite adequate for frequencies below about 1000 GHz.

Thermal noise, then, establishes a fundamental lower signal level limit in processing and measurement of signals (amplification, for example). You can define the minimum signal that can be processed as that signal level which results in a signal-to-noise ratio of unity. (Remember, this is the signal level required to *just* be able to discern that a signal is present.) Any signal in a system that's smaller than the thermal noise at that point in the system will be lost in the noise. You can examine a simple resistor to find the signal level that results in a signal-to-thermal noise ratio of unity with this resistor as the source resistance. Consider a resistor of value R_S at an absolute temperature T_S. Let the noise bandwidth of the "viewing" instrument of the source (amplifier noise bandwidth or measurement system noise bandwidth) be BW_n. The thermal noise power available from this resistor to a matched load is given by **Equation 1**. For a simple resistor, the matched load is a resistance of equal value — in this case, R_L. **Figure 1** shows a "warm" resistor, your source, loaded into a "noiseless" matched load. Actually, there's no such practical noiseless load. It's used here as a mathematical tool to examine the noise processes.

Signal bandwidth

The noise power available to the matched load from the warm source resistance, given by **Equation 1**, is also the *signal* power level at which a power signal-to-thermal noise ratio of unity would be obtained with "perfectly noiseless processing." Any signal lower than that level will be below the noise and essentially unrecoverable. You can do only two things to reduce the fundamental noise limit in this simple system — reduce the temperature of the source or reduce the noise bandwidth. In practice, the source temperature is usually determined by processes beyond your control. The signal bandwidth, and therefore the noise bandwidth, is generally selected to pass the required spectral contents of the signal being processed. This means that the thermal noise can't be reduced below the fundamental limit given in **Equation 1**. However, you can draw one important conclusion. In low noise designs, it's important to incorporate a signal bandwidth that's just wide enough to pass the signal of interest with the needed fidelity. If a wider bandwidth is used, no additional signal improvement is gained but additional noise is added. For example, if you want to pass a DC-1 MHz signal, and use a bandwidth of DC-10 MHz, the noise power will be a factor of 10 dB higher than if you used the minimum bandwidth of 1 MHz.

FIGURE 2



Thermal noise source circuit models.

Warm resistor noise model

Johnson's fundamental noise theory lets you find the value of noise power available at the matched load in **Figure 1**. Using this information, you can derive a noise model of the warm resistor. The resistive noise source may be modeled as a voltage source in series with its characteristic resistance (a Thevenin model) or as a current source in parallel with its characteristic resistance (a Norton model). These are exactly equivalent in linear systems, but the voltage source model is generally the model used. These two circuit models are shown in **Figure 2**. Now you know the noise power delivered to the noiseless matched load from **Equation 1**. The voltage that must be impressed across that load to produce that power is easily computed:

$$P_r = E_r^2/R = k\ T\ BW_n$$

so:

$$E_r = \sqrt{P_r R} = \sqrt{k\ T\ BW_n\ R} \quad (rms) \qquad \textbf{(3)}$$

A similar computation may be performed for the current in the load.

$$P_r = I_r^2\ R = k\ T\ BW_n$$

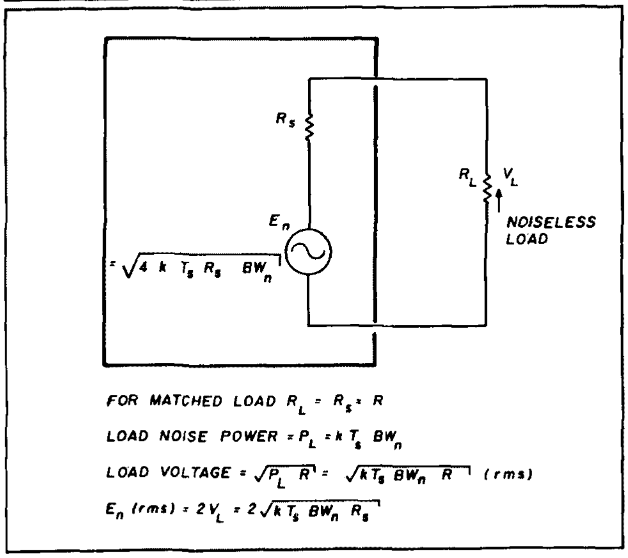
$$I_r = \sqrt{k\ T\ BW_n/R} \quad (rms) \qquad \textbf{(4)}$$

These results are the noise voltage and current, respectively, delivered to the matched load from the warm source. From these you can compute the equivalent voltage and current sources in **Figure 2**. The computation steps for the Thevenin source model are shown in **Figure 3**. Because the source and load resistances are equal, the equivalent source voltage is simply twice the voltage delivered to the load (one-half of the source voltage is dropped across the source resistance and one-half across the load resistance). The same is true of the equivalent current source in terms of current, so:

$$E_n = 2\sqrt{k\ T\ BW_n\ R} \quad (rms) \qquad \textbf{(5)}$$

$$I_n = 2\sqrt{k\ T\ BW_n/R} \quad (rms) \qquad \textbf{(6)}$$

FIGURE 3



Thermal source Thevenin potential computation.

Matched and unmatched loads

The noise model values were derived from Johnson's findings for matched loads. However, these are true linear models and represent true resistive sources. They may be used accurately in any application independent of the load they are driving. In cases where a matched load is being driven, the maximum power is delivered and is that given by **Equation 1**. For unmatched loads, the delivered power will be less than the matched case and must be computed directly from the specific circuit element values.

Limiting value of source voltage

The noise voltage given by **Equation 5** is the minimum possible noise voltage that can be achieved for a resistor at the specified temperature and noise bandwidth. If this resistance is the source resistance of some device, it is a limiting value of source voltage. Consider that you are attempting to amplify small signals with a 1-MHz signal bandwidth from a 50-ohm room temperature source. Using

Equation 5, the equivalent thermal noise voltage of the resistor may be computed:

$$E_r(50\ ohms) =$$

$$2\sqrt{(1.38\times10^{-23}\ W\text{-}s/^\circ K)\ (290^\circ K)\ (\pi/2\ MHz)\ (50\ ohms)}$$

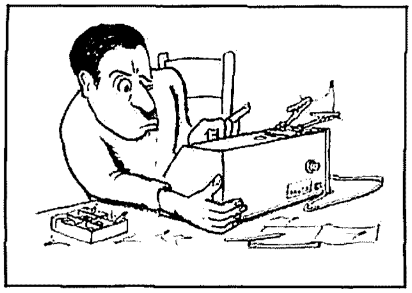
$$= 1.12\ \mu V(rms)$$

Thus, the signal for which the signal-to-noise ratio at the source is unity in this example is 1.12 μV rms. If a unity signal-to-noise ratio is defined as the point at which the signal is just lost in the noise, this thermal noise level is the limiting value of signal level — or about 1 μV in this example. This is the best that can be achieved in minimizing noise. Even with a perfect, noiseless amplifier with a noise figure of 0 dB or noise temperature of zero degrees, this would be the limiting noise. If the signal is below this limiting value, you can't process it out of the noise — even with a perfect, noiseless amplifier.

Part 2 will discuss noise figure phenomena and amplifier equivalent noise temperature.

1. C.D. Motchenbacher and F.C. Fitchen, *Low-Noise Electronic Design*, Wiley, New York, 1973.
2. S.P. Parker, *McGraw-Hill Dictionary of Scientific and Technical Terms*, Third Edition, McGraw-Hill, New York, 1984.

Practically Speaking



Joseph J. Carr, K4IPV

REVIEW OF A LOW COST COMMERCIAL SPECTRUM ANALYZER

One of the most popular topics I've covered in this column concerned a build-it-yourself spectrum analyzer. A lot of readers put together those kits and modified them considerably. I've published some of my modifications, and those of a few readers, in this column over the past two years. A few months ago I told you that a small manufacturer of low cost spectrum analyzers was going to make one of their products available for review in this column. Judging from my mail, that announcement interested a lot of you. Well, Penntek Instruments (14 Peace Drive, Lewistown, Pennsylvania 17044 (717)248-2507) came through with a loaner instrument, so here's the promised review.

What is a spectrum analyzer?

For those who may have missed the earlier articles, I'll review briefly what a spectrum analyzer is and how one can be used in communications work. The spectrum analyzer is a special form of swept superheterodyne receiver with a frequency domain output (*amplitude versus frequency*) instead of the time domain output usually found on oscilloscopes. To display the output of a spectrum analyzer you must use an X-Y oscilloscope, rather than the standard Y-time type. Most common oscilloscopes can be used in X-Y mode if they have a horizontal input, or are dual beam models in which an X-Y mode is provided on the vertical function selector.

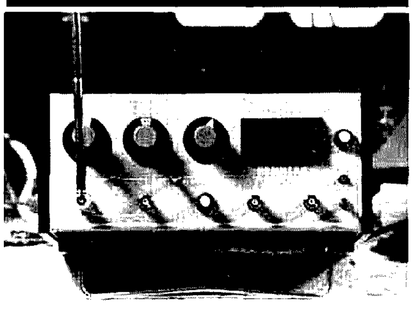
An internal sawtooth waveform causes the receiver tuning to sweep from one end of the band to the other. The same sawtooth waveform is also directed to the X-input (horizontal) of the oscilloscope, where it's used to sweep the CRT beam from left to right. As the receiver and scope beam sweep through the range, the detected output is fed to the Y-input (vertical) of the oscilloscope. This causes an amplitude-versus-time display to appear.

Penntek Instruments' Model SA-500E spectrum analyzer (**Photo A**) converts any oscilloscope that has either an X-Y mode or allows access to the horizontal amplifiers (which is the method used on older oscilloscopes). The display outputs from the SA-500E are scaled to 1 volt/division; the X-output is ±5 volts and the Y-output is ±4 volts. The horizontal sweep speed (X-output) is variable from 1 to 40 Hz.

Penntek Model SA-500E spectrum analyzer

The SA-500E has a frequency range of 0 to 550 MHz, with a digital dial accuracy of 2 percent of full scale. This instrument is basically a swept triple conversion superheterodyne receiver. There are twelve switch-selected sweep (or "span") widths from 50 MHz/division down to 20 kHz/division in the popular 5:2:1 sequence ratio. When the SA-500E is set for the 50-MHz/division span width, the entire spectrum from 0 to 550 MHz is displayed on the oscilloscope. There's also a 0-MHz/division sweep width, in which the SA-500E operates as a manually tuned receiver or wave analyzer. The frequency resolution or IF bandwidth has two options — wide (200 kHz) and narrow (10 kHz). Bandwidth selection is coupled to the sweep width control. The input sensitivity is better than 2 μV and the receiver has a 70-dB dynamic range. An RF input attenuator allows selection of up to 70-dB attenuation in 10-dB steps, while an IF attenuator offers a variable range of 0 to 22 dB. Allowable input levels are 0 to +13 dBm RF, 1 volt AC, and 0 volts DC. There are three IF frequencies used in the SA-500E: 700 MHz, 62 MHz, and 10.7 MHz. The SA-500E offers two crystal marker frequencies to help calibrate the dial, 5 and 50 MHz.

PHOTO A



Penntek SA-500E spectrum analyzer.

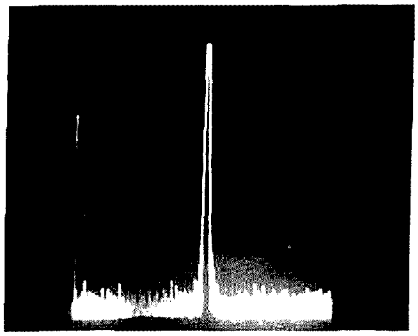
The Penntek SA-500E is small and doesn't take up a lot of space on the bench. Its dimensions are 10-3/8 × 11-3/4 × 5 inches and it weighs under 3 pounds. The cabinet is gray and black.

Using the SA-500E spectrum analyzer

The SA-500E is intended for operation with an external oscilloscope. The X and Y outputs of the spectrum analyzer must be connected to the X (horizontal) and Y (vertical) inputs of the oscilloscope. The scope input attenuators must be set to 1 volt/division or 2 volts/division, whichever gives the best display. You then turn the center frequency control to the center of the band of interest and set the span control to the desired sweep width.

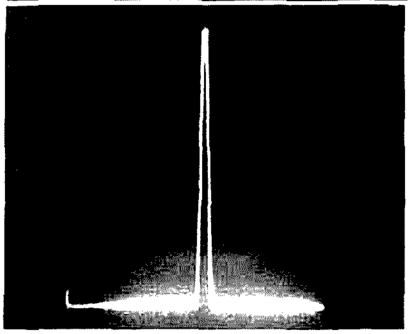
Photos B and C show the SA-500E display that appeared when I connected my Measurements Model 80 signal generator to the input of the spectrum analyzer. The signal generator's output frequency was approximately 100 MHz, plus or minus the

PHOTO B



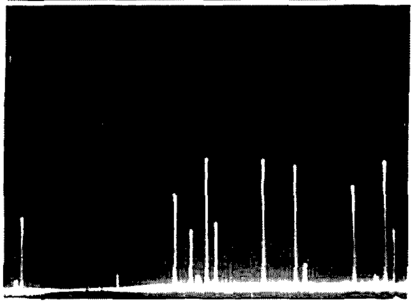
Signal generator output signal without video filter.

PHOTO C



Signal generator output signal with video filter.

PHOTO D



Display of 2-meter band signals near Washington, D.C.

30-year old calibration. The signal in Photo B shows the display that resulted when I selected no filtering on the SA-500E. When I selected the video filter, the baseline noise disappeared and the trace in Photo C resulted. This display was taken with the span control set to 0.05 MHz/division, so the trace represents a 5-MHz slice of spectrum centered on 100 MHz.

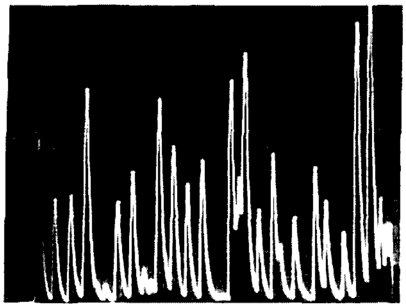
After I played with the process of displaying a signal from my signal generator on the scope, I connected the RF input of the SA-500E to the 2-meter ground plane antenna that serves as my access to the local repeater world. Photo D shows a 1-MHz/division slice of spectrum centered on 146 MHz. A large number of local repeaters, mobiles, and fixed stations showed up on the display. The spikes representing these signals appeared and disappeared as the stations went on and off the air.

Monitoring the 2-meter band on my old Kenwood transceiver revealed that the centermost spike was from the 146.91-MHz repeater in Fairfax, Virginia. That spike disappeared every time the repeater went off the air. Interestingly enough, most of the time I could also see a much smaller spike ride up and down with the repeater spike; I believe this represented mobile rigs accessing the repeater. Because the repeater receiver is a lot more sensitive than the spectrum analyzer and installed at an elevated location, it could hear more mobile signals than the spectrum analyzer. This accounts for the fact that the mobiles didn't always show up on the scope screen. Only those mobiles that were relatively local would show up.

I tuned the SA-500E center frequency control to the middle of the FM broadcast band (88 to 108 MHz) and looked at the scope display. Photo E shows the FM band viewed from my location in Virginia, near Washington, DC The large spike on the right is from my Model 80 signal generator — which I'd failed to turn off before retuning the spectrum analyzer.

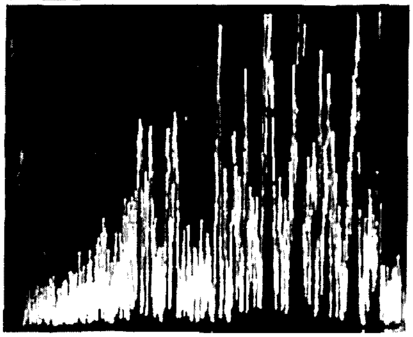
No one who looks at Photo E can doubt that the FM broadcast band is crowded in my area, but wait until you see the HF spectrum. No denizen of 20 meters will deny that HF is overrun with signals; Photo F demonstrates that lamentable fact rather conclusively. This trace represents the HF spectrum centered on 14 MHz, with a sweep span that lets you look about 5 MHz above and below that frequency. The big center group of spikes represents the 20-meter band, while those to the right seem to show a fairly active international broadcasting band. I'm sure that the WWV time and frequency standard station is somewhere in that mess!

PHOTO E



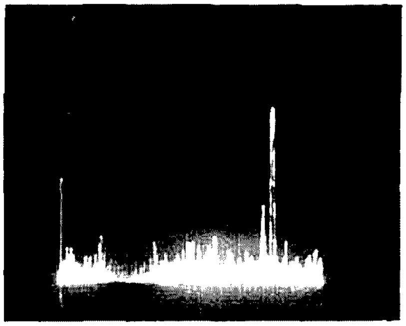
Display of the FM broadcast band signals (88 to 108 MHz) near Washington, D.C.

PHOTO F



High frequency (HF) bands on the spectrum analyzer shows why 20-meters is so maddening.

PHOTO G



"Spur" turned out to be the local oscillator of an FM broadcast receiver sitting next to the spectrum analyzer.

Photo G perplexed me a little bit at first. Initially I thought the signal was a

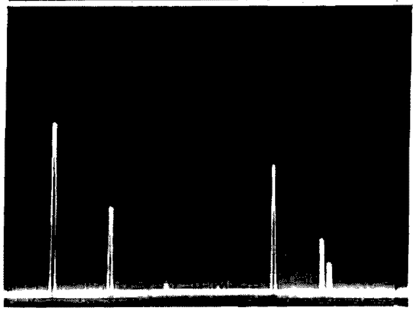
spurious reponse of the SA-500E (most spectrum analyzers have at least a few). But the signal didn't seem to behave like a spur. Then I found the problem. I listen to radio while working at the word processor and at my workbench. Those who know me personally can attest to my love of (blush) bluegrass music — the more traditional the better. I had the radio tuned to 88.5 MHz (WAMU in Washington, DC) for the Saturday bluegrass show. The SA-500E antenna was disconnected and I was using only a short, shielded scope probe to supply input signal from my signal generator output. But even when the signal generator was turned off, this little signal persisted.

On a hunch, gleaned from an awful troubleshooting experience many years ago, I turned off the FM broadcast receiver on the workbench. I'm sure you can guess what happened. The "spur" disappeared. The signal displayed in Photo G is the radiated local oscillator signal from my FM receiver. It was on a frequency of 88.5 + 10.7 MHz, or 99.2 MHz. The rest of the FM band didn't show up because of the short "antenna" represented by the partially shielded scope probe.

As I mentioned in November's column, I saw this same problem more than ten years ago when I worked at a major university medical center. Nurses in the post-coronary care unit complained that one patient's electrocardiograph (EKG) signal was showing up on two channels of the radio telemetry display. The problem turned out to be an unshielded FM receiver located too close to the telemetry receiver's local antenna. The receiver was "intermodding" (for lack of a better term) with the other signals in the system, forcing the weak EKG radio signal onto a different frequency. Penntek tells me that a lot of hospital biomedical equipment shops have bought their spectrum analyzer. I wish I'd had one that night a decade ago when I was trying to figure out the solution to that perplexing problem. Now it's easy to understand why airlines don't like FM radios on board. The local oscillator of FM receivers falls into the aviation band when the FM radio is tuned to the high end of the FM broadcast band!

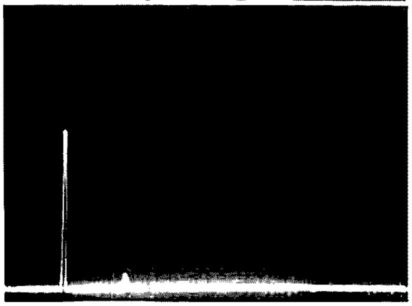
While it's fun to look at the spectrum through an instrument like the Penntek SA-500E, there must be a practical use for it before most of us will invest in one. There are several uses of interest to the

PHOTO H



Output spectrum of a frequency tripler circuit when output had previously been peaked.

PHOTO I



Output spectrum of a frequency tripler circuit after tuning with aid of spectrum analyzer.

Amateur. In addition to identifying intermod problems, it's also possible to use the SA-500E to tune transmitters and other RF devices. Amateurs typically tune RF circuits for maximum output or to null some interfering signal. The SA-500E offers a means for doing this job easily.

Photos H and I show the responses of a VHF oscillator that's part of a converter project I'm building (see Figure 1). The local oscillator (LO) frequency is generated in a crystal oscillator at 43 MHz and then tripled to 129 MHz. The tripler is a diode multiplier circuit followed by a tuned RF amplifier at the 129-MHz output frequency (a circuit derived from, but not exactly like, an old *ARRL Handbook* project).

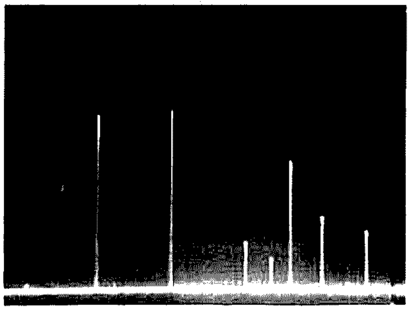
There are really two issues involved when you're tuning the desired output of the frequency multiplier. One concern is to maximize the output signal so it can be used to drive the mixer. The other is to minimize the harmonics and other spurious responses. These signals can interfere with the proper receiver operation.

In Photo H the leftmost spike is the 129-MHz output signal, while its second harmonic (258 MHz) is the third spike from the left (second highest). There are three other spikes in the spectrum. I'm not sure where they came from, but they weren't supposed to be there. The spacing on the scope display suggests that they might be other harmonics of 43 MHz, or mixing products of the 43 MHz plus its other harmonics.

Photo I shows what results when you tune the multiplier (C1 in Figure 1) and the output of the amplifier (C2) properly. The main signal at 129 MHz lost a little amplitude, but the harmonics and/or spurs disappeared. This is what it's supposed to look like.

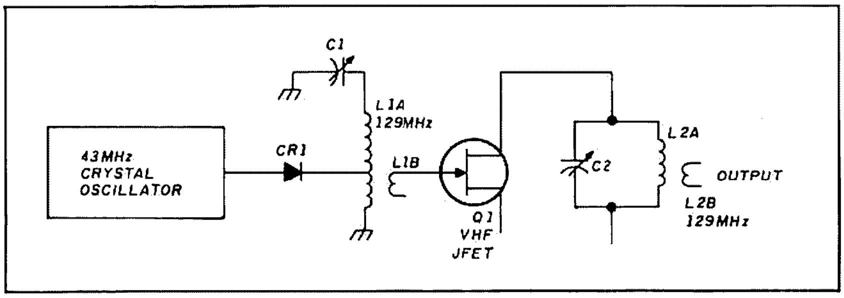
Photo J shows another application

PHOTO J



Bandsweeper mixer output signal when one oscillator was acting "funny."

FIGURE 1



VHF tripler used to generate the signals in *Photo G*.

of the spectrum analyzer in adjusting Amateur projects. I'm building a bandsweeper — a sweep signal generator that will cover an entire HF Amateur band at one setting. The project uses two low band VHF voltage-controlled oscillators (VCO) heterodyned together in a Mini-Circuits balanced mixer. One will be fixed at a center frequency and then swept with a sawtooth (the digital sawtooth generator reported in this column earlier); the other will be fixed tuned to a frequency that heterodynes the sweep oscillator to the desired Amateur band. Both oscillators are adaptations of an oscillator circuit that was used in some commercial test equipment.

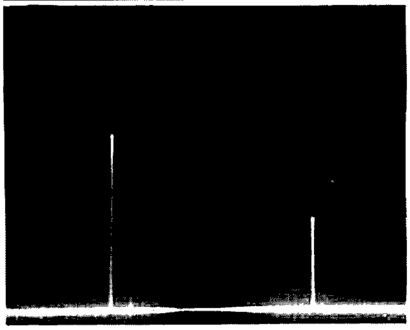
When I looked at the 15-meter band amplitude versus time display on my oscilloscope, there was a lot of disconcerting distortion of the signal that I believed should be nearly sinusoidal. Looking at the 15-meter output of the project on the SA-500E (Photo J) showed the cause of the distortion. There were a large number of mixer products and harmonics present.

I discovered two problems that needed attention. First, the oscillator feedback control (C1 in **Figure 2**) was improperly adjusted. Second, I needed either a low pass or a notch filter on the mixer output. Photo K shows what happened when I adjusted the feedback control to a point between its initial adjustment and the point where oscillation ceased. (If anyone can propose a reason for this spectrum, I'd appreciate it. Just write to me at the address at the end of the column.) The trace in Photo K shows that the mixer product is still in the output, so I'll add a filter before I publish the circuit in this column.

As you can see, there are several possible Amateur applications for the spectrum analyzer. You can use it to adjust the tuning of projects in progress. You can also use it to test transmitters for harmonic radiation. Federal Communications Commission rules, common sense, and plain old decency require that we suppress harmonic emissions as much as possible. Whether the equipment is homebrew or store bought, it's the transmitter's licensee who's responsible for the correct operation of the unit. The spectrum analyzer helps because it lets you test transmitters for spurious emissions.

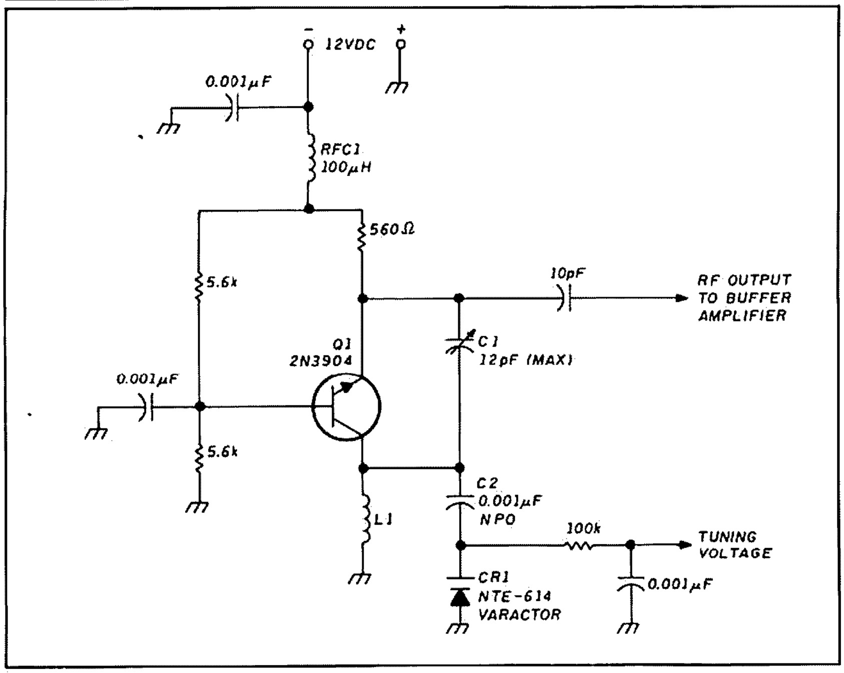
Finally, you can use the spectrum analyzer to locate sources of RF that are interfering with your own operations. It may also be helpful in locating TVI sources. It's quite possible for Amateurs to be blamed for TVI that isn't their fault — neighbors can be like that. (In my Novice days, an irate neighbor blamed me for TVI even when I was asleep, and before my first Heathkit transmitter arrived in the mail! I guess it was those 500-THz Martians who were using my receiver dipole. Seriously, though, a spectrum analyzer helps make such hidden signal hunts a lot easier.) I'm sure many other applications will surface once a few Amateurs start using this instrument.

PHOTO K



Bandsweeper mixer output signal when the feedback level was readjusted.

FIGURE 2



Oscillator used in the bandsweeper project.

Conclusion

The Penntek SA-500E spectrum analyzer is a viable and useful tool for Amateurs. It costs about the same as a decent HF SSB rig ($1,495), so it may be a little pricey for many people. But if you're into serious technical work, or if your club would consider buying an instrument for everyone to use, you should consider this little analyzer. Penntek will be happy to send you a spec sheet and answer any reasonable technical questions.

NOTE: Several readers have written to me, or approached me at hamfests, to suggest a panadapter project. A panadapter is like a spectrum analyzer, but tends to be fixed tuned to the IF frequency of a ham receiver and sweeps sufficiently to cover the entire band to which the receiver is tuned. Some receiver and transceiver manufacturers make these units as add-on adapters. Does anyone have any ideas for an Amateur panadapter centered on either 8.83 or 9 MHz? If so, then drop me a line, or better yet, write an article yourself and submit it to *Ham Radio*!

Joe Carr, K4IPV, can be reached at POB 1099, Falls Church, Virginia 22041. He would appreciate your questions and recommendations for this column.

NEW PRODUCTS

New Quad Clip™ Test Adapter Kit

Pomona Electronics offers all six of its current family of Quad Clip test adapters in kit form, packaged in a convenient carrying case. Units are designed to fit PCC or PLCC ICs with "J" leads. The kit, Model 5515, includes one each of Pomona's 20, 28, 44, 52, 68, and 84-pin Quad Clip test adapters.

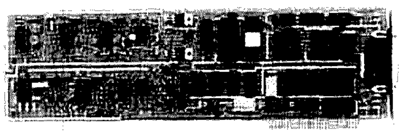
Quad Clip adapters use Pomona's patented snap ring locking system to mount securely on the IC under test. Output signals are delivered by staggered 0.025-inch square gold-plated pins.

Model 5515 is described and illustrated in the 1989 Pomona Electronics general catalog. Copies are available without charge from ITT Pomona Electronics, 1500 E. Ninth Street, Pomona, California 91766. Phone: (714)623-3463. FAX: (714)629-3317.

Circle #301 on Reader Service Card.

HAL PC-AMTOR

PC-AMTOR is a new HAL product designed specifically for Radio Amateur AMTOR, RTTY, and Morse code operation. The PC-AMTOR circuit board plugs directly into an IBM-compatible PC and includes user-friendly terminal software on a 5.25" diskette.



PC-AMTOR features improved AMTOR performance and simplified use. Both CCIR-476 and the new CCIR-625 AMTOR protocol are included. HAL software uses pull-down menus. Modes and features may be changed either via menu or by using "expert" key letter commands.

PC-AMTOR also has standard Baudot and ASCII RTTY, and an improved Morse code send/receive algorithm. A new auto receive mode takes the guesswork out of monitoring.

PC-AMTOR includes a new "host mode" serial I/O port which lets you use PC-based APLink or mailbox programs — programs that support only PC serial I/O modem connections.

PC-AMTOR is a full length PC-compatible card. It requires a standard PC-XT or PC-AT with 640K of RAM and a minimum of one 360K floppy disk drive. Monochrome, CGA, and EGA video are also supported. PC-AMTOR does *not* require standard PC serial or parallel I/O interrupts or addresses.

PC-AMTOR is priced at $395, including software. The model number is PCI-3000. A new SPECTRA-TUNE tuning indicator, the SPT2, with integral cable expansion and simple connections to other devices, will soon be available.

For details contact HAL Communications Corporation, PO Box 365, Urbana, Illinois 61801. Phone: (217)367-7373. FAX: (217)367-1701.

Circle #302 on Reader Service Card.

"Fast Code" Test Prep Tapes for ARRL Exams

Gordon West announces new Morse code test preparation tapes designed specifically for the new ARRL "Fast Code" CW examinations. Code

characters are generated at 18 wpm and spaced for the 5-wpm ARRL Novice and the 13-wpm ARRL General tests.

Radio School tapes are now distributed by the Radio Amateur Callbook, Inc., PO Box 247, Lake Bluff, Illinois 60044. Phone: (312)234-6600. They are also available through most major Amateur Radio dealers.

Circle #303 on Reader Service Card.

ARMOR-FLEX™ Line of Portable Antennas

The new ARMOR-FLEX™ line of portable replacement antennas is now available from The Antenna Specialists Company. There are more than 100 different PVC-dipped or polyurethane molded models for low band (30 to 50 MHz), VHF (141 to 174 MHz), UHF (406 to 512 MHz), and trunking and cellular (800 to 900 MHz). A new brochure, including a handy selection guide, is available on request. The selection guide provides a cross-reference chart for quick, accurate antenna selection by radio type, connector requirements, and frequency.

For more information, contact The Antenna Specialists Company, 30500 Bruce Industrial Parkway, Cleveland, Ohio 44139-3996. Phone: (216)349-8400. FAX: (216)349-8407.

Circle #304 on Reader Service Card.

New Voice Digitizer Has Added Features

QRZ Industries of Piedmont, South Carolina announces the uVB-1 natural voice digitizer. The uVB-1 is a single message voice digitizer which can store up to 8 seconds of voice (expandable to 32 seconds with built-in memory expansion port). The unit has two auto repeat modes. One has rapid adjustable repeat for contesting (0 to 20 seconds); the other has a programmable long repeat mode for repeater IDs (up to 20 minutes). A built-in monitor circuit, capable of driving headphones or a speaker, combines digitized and live audio with an auxiliary audio input. There are separate adjustments for balancing live voice and digitized audio, and setting auxiliary audio gain, 600-ohm ouput level, and monitor output level. To program your voice into the uVB-1 simply speak into the microphone while the unit is in record mode.

The uVB-1X is similar to the uVB-1, but it uses CMOS static RAM for extremely low power consumption and provides on-board memory lithium battery backup. A special power-down circuit ensures that memory contents are maintained when power is lost.

Model VB-8A

The VB-8A is the next generation replacement for the popular VB-8. Model VB-8A has the basic features of the VB-8, along with a number of improvements and enhancements. Featuring eight soft-sectored messages, the VB-8A is capable of recording up to 100 seconds of digitized speech. The audio quality has been improved by increasing the digitizing rate to 40 kHz and adding 14 poles of audio filtering. A built-in monitor amplifier is standard and an automatic serial number inserter (which uses the operator's own voice) has been added.

(continued on page 88)

JUNKBOX VARIABLE CRYSTAL OSCILLATOR

By Charlie Tiemeyer, W3RMD, RD 2, Box 427-C, Chestertown, Maryland 21620

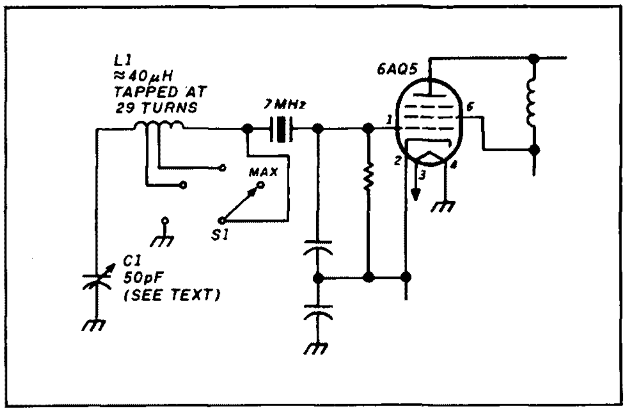
I built an 80/40 Meter Junkbox rig and needed a compatible crystal oscillator. I've come up with one that lets you "rubberize," or move your crystal frequency several kilohertz above and below the fundamental. Most of my "rocks" are the FT-243 type, but I'm sure any type you may have can also be shifted.

The rig mentioned above uses a 6AQ5 oscillator tube, but any other oscillator tube should be adequate. My well-modified Globe Chief Model 90A uses a 6AG7 oscillator tube and this VXO works in it equally well. The only requirement is that your oscillator tube have a grounded grid through the crystal.

VXO circuit

Figure 1 shows the schematic diagram of the Junkbox VXO. Note that it's only necessary to install or insert coil L1 and variable capacitor C1 in series between the crystal and ground. If you have room, it would be more convenient to add these components to your rig. I built my oscillator in a minibox approximately 4" wide × 3" deep × 2-1/2" high for outboard use. Before this, however, I did install a VXO in the transmitter itself.

FIGURE 1



Schematic of the Junkbox VXO. The 6AQ5 oscillator circuit is described in Amateur handbooks.

If you don't have a minibox, you can get small metal cabinets in various sizes from Radio Shack. Two possibilities are their 3-1/2" × 2-1/8" × 4" cabinet (catalog no. 270-251) and their 4" × 2-3/8" × 6" cabinet (catalog no. 270-252). With a little ingenuity, you could probably use an empty coffee can, or something else of reasonable size. I suggest strongly that you use the smallest possible enclosure so that all leads are as short as possible. This will minimize unnecessary capacitive reactance. Make sure your enclosure is well shielded to reduce TVI.

Some years ago I built an inboard VXO for my rigs using a slug-tuned coil. A company called Calectro once made them in various ratings. I recall using one to cover 5 to 25 μH. It worked pretty well, but is no longer available. The amount of frequency shift wasn't as great with these slug-tuned coils as it is with the coil used in this circuit.

The LC circuit

Basically I use 7-MHz crystals, but I can adjust the plate circuit for output at 7, 14, or 21 MHz. I can also use 3.5-MHz crystals and swing the frequency — although not as much as I can with the 7-MHz crystals. L1, which is about 40 μH, works in series with C1 and the ground side of the crystal to swing the frequency. You can adjust the circuit to furnish either capacitive or inductive reactance to the crystal with these two components in a series-resonant circuit. This means C1 can tune from slightly above to considerably below the crystal's normal frequency.

About 35 pF is all that's normally necessary for full range with enough inductance to achieve the desired frequency shift. I didn't have a 35-pF variable capacitor, so I'm using a 50-pF variable. It works okay, but uses only the upper 30 to 35 pF. If you close or fully mesh the plates of a capacitor that's more than 35 pF, you'll notice that your crystals won't oscillate. However, the circuit will oscillate with a different tap on L1 and also with 80-meter crystals. A little experimentation is required.

The inductance of L1 is around 40 μH. To determine this inductance, refer to the chapter "Electrical Laws and Circuits" in the *ARRL Radio Amateur's Handbook*, under the heading "Calculating Inductance." I happened to have an old Globe Chief tank coil 1 inch in diameter, using about no. 18 enameled wire, close wound. I use 58 turns, which approximates 40 μH, and a tap at 29 turns — the half-way point.

Using coil taps

You may have a larger diameter coil or form. This might be a little better, but it requires fewer turns to achieve 40 μH. It may be wise to have at least two taps, perhaps several, which should be close to the lower end of L1.

By using taps, you can reduce the shift or use 10-MHz crystals for the 30-meter band. The inductance is critical; too little gives insufficient shift, while too much produces a too fast tuning rate. It's best to use just enough inductance to shift to the desired frequency with the entire 30 to 35 pF of C1. Of course, as I mentioned before, if you have more capacitance in your variable capacitor you'll use only the upper 30 to 35 pF.

My L1 coil has just *one* tap because I use a bat-handle toggle switch to short half the coil. I found that by using this method with my 50-pF variable capacitor, I can tune a little higher or lower than the crystal frequency — depending on the switching arrangement used. You must experiment to achieve your desired results, but about 40 μH of inductance is necessary.

There you have it, friends of the ether. I'm sure you will have satisfactory results with this VXO. If you have only a handful of crystals, you can change frequency up or down without being completely "rock bound." I find this method gives crystal stability with VFO capability.

You may find a few things to enhance this circuit. I'd appreciate any comments you have to offer. hr

HT MOBILE COMPANION

Handheld accessory uses just three circuit boards

By Peter A. Lovelock, K6JM, 1330 California Avenue, Santa Monica, California 90403

Handheld VHF and UHF transceivers (HTs or handy talkies) are popular with Amateur operators. Many use HTs for mobile operation; this gives you two radios for the price of one. However, using an HT in this mode has disadvantages. Low audio comes into a tiny speaker and may be drowned out by traffic noise before it reaches your ears. You might also experience battery discharge in the middle of a QSO.

The Mobile Companion alleviates these difficulties, and does so in a compact unit. One interconnecting cable to the HT provides the following features:

- Regulated 10.5-volts DC at 1 A to power the HT and save the battery.
- A constant voltage charger, which quickly replenishes a dead battery pack.
- A receiver audio booster, which delivers 2 or 8 watts to an external speaker.
- An optional PA system for emergency use and monitoring the receiver when you're not in the car.

A full schematic is shown in **Figure 1**. The battery charger and 8-watt amplifier (which doubles for receiver audio boost and PA operation) are included.

Not everyone will want all the features. I used three small pc boards for the audio amplifier and battery charger — which offer simple, flexible choices.

Some HTs have a jack accepting up to 16 volts DC directly from the car battery. This doesn't protect against transient switching spikes or auto regulator failure, which may place a harmful voltage on the HT. The LM7805 regulator provides damage insurance; it helps relieve filter alternator whine and other noise. The regulator is set for 10.6 volts DC (the most common supply for HTs) and maintains regulation with the car battery to 12 volts DC. This means a sacrifice in RF output power for 5-watt RF HTs using 12-volt battery packs.

I needed a small under-the-hood PA speaker for use with our local emergency team. I replaced the HT's 2-watt amplifier board with an 8-watt commercial module (see parts list). Two miniature relays switch the amplifier input and output for radio or PA mode.

Functional description

SW1 and SW2 (DPDT slide switches) select the radio and PA modes, as shown in **Table 1** and **Figure 1**.

Setting SW1A in the R position connects the mic to the HT mic input for normal radio operation; in the P position it connects the mic to preamp Q1 for PA mode. Putting SW1B in the P position turns on flashing LED CR1 and energizes K2 to connect J1 (PA speaker) to U2 amplifier output. Coil K1 is also supplied voltage, but SW2A in the R position keeps K1 de-energized and the amplifier input connected to R8 (radio gain control), for radio monitoring through the PA speaker. Placing SW2A in the P position allows K1 to energize through SW2B, connecting the amplifier input to R10 (PA gain control), and operating the PA from the mic. In the P position SW2B causes LED CR2 to light, showing PA mode.

TABLE 1

Mode select arrangement

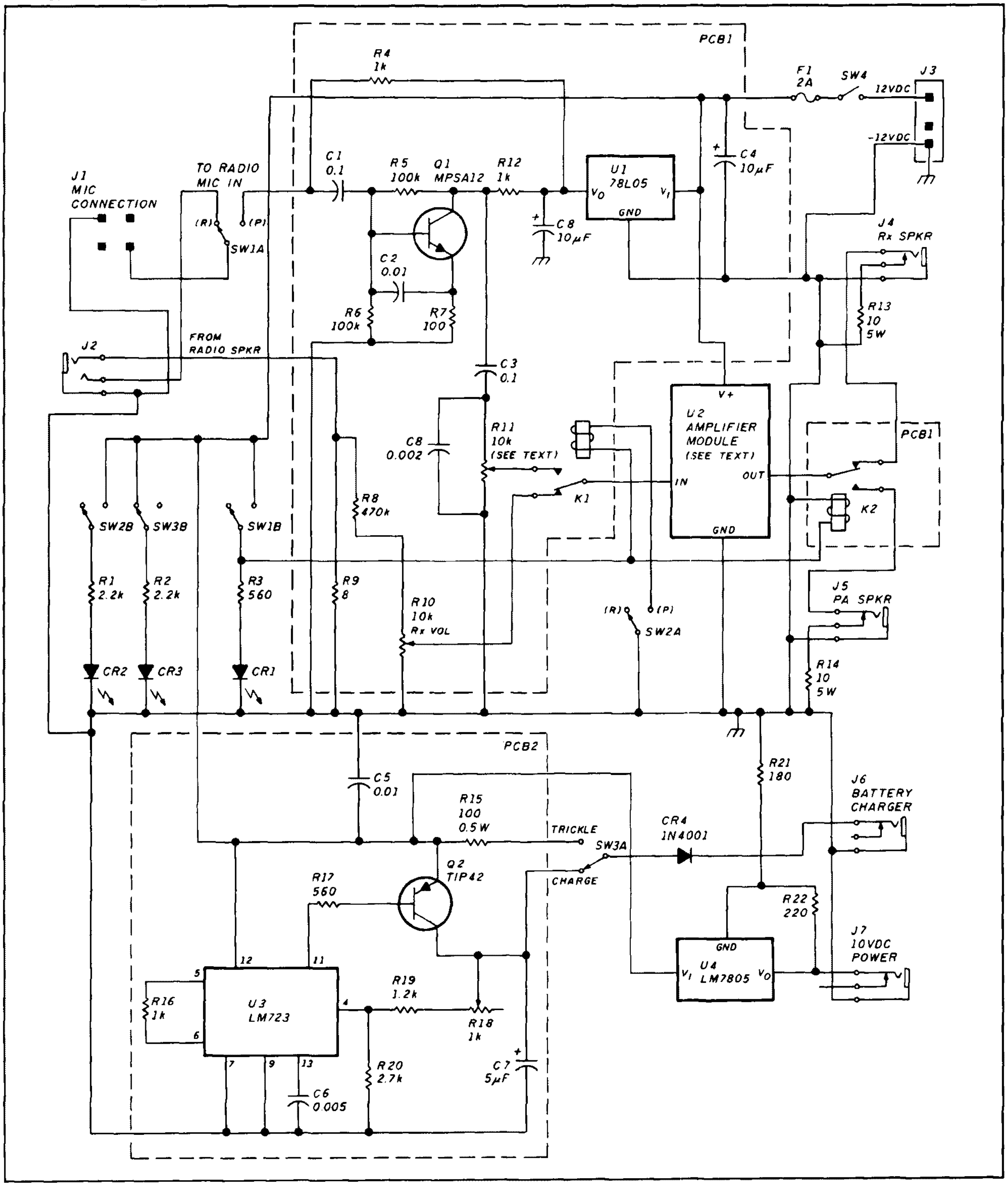
SW1	SW2	Mode
Radio	PA	Radio
Radio	RX	Radio
PA	PA	Mic/PA
PA	RX	RX/PA

Preamp Q1 provides gain for the mic to drive the 8-watt amplifier to full output. It's configured for a high-impedance dynamic or Electret™ (capacitive) mic. *R4, a 1-k resistor, is required only with an Electret mic and should be excluded when you're using a dynamic or ceramic mic.* VCC 5 volts for the preamp is supplied by U1, a 78L05 regulator (which also helps isolate car-generated noise from the mic line).

The charger circuit was originally developed by Joe Moell, WØOV. It comprises Q2 — a low-saturation PNP pass transistor TIP42A — controlled by U3, a μA723 voltage regulator. *Do not* substitute another transistor for the TIP42A. The charger circuit is an adjustable constant-voltage type.

A fully discharged battery pack will draw high current (about 750 mA) at first, tapering down to approximately 25 mA as the battery pack is fully charged. SW3A selects full charge or a trickle charge of 10 mA; SW3B controls CR3 to show when the full-charge status is on.

FIGURE 1



Schematic of the Mobile Companion.

PARTS LIST

Reference	Description	Radio Shack	Quantity
PC BOARD 1, CONTROLLER			
Q1	*Transistor MPSA12*		*1*
U1	*Voltage regulator 78L05*		*1*
K1,K2	*Relay, micro, SPDT, 12-volt coil*	*275-241*	*2*
R10	*Trimmer potentiometer, 10 k*	*271-335*	*1*
R4-R12	*Resistors, 1/4 watt, composition*		*9*
C1,C3	*Capacitor, ceramic, 0.1μF, 16 volts DC*		*2*
C2	*Capacitor, ceramic, 0.01μF, 16 volts DC*		*1*
C4,C9	*Capacitor, 10μF, 25 volts DC*		*2*
C8	*Capacitor, ceramic, 0.002μF*		*1*
PC BOARD 2, CHARGER			
Q2	*Transistor, PNP power, TIP42*		*1*
U3	*IC voltage regulator, LM723*	*276-1740*	*1*
R16,R17, R19,R20	*Resistor, 1/4 watt, composition*		*4*
R18	*Trimmer potentiometer, 1 k*	*271-333*	*1*
R15	*Resistor, 100 ohms, 1/2 watt*		*1*
C5	*Capacitor, ceramic, 0.01μF, 16 volts DC*		*1*
C6	*Capacitor, ceramic, 0.005μF, 16 volts DC*		*1*
C7	*Capacitor, tantalum, 5μF, 16 volts DC*		*1*
	Insulated mounting kit for Q2	*276-1373*	*1*
PC BOARD 3, OPTIONAL 2-WATT AUDIO AMPLIFIER			
U4	*IC amplifier, LM380*		*1*
C1	*Capacitor, ceramic, 120 pF, 10 volts DC*		*1*
C2	*Capacitor, electrolytic, 47μF 10 volts DC*		*1*
C3,C4	*Capacitor, ceramic, 0.1μF, 16 volts DC*		*2*
C5	*Capacitor, electrolytic, 470μF, 16 volts DC*		*1*
R1	*Resistor 2.7 ohms, 1/2 watt, composition*		*1*
CASE COMPONENTS			
U4	IC voltage regulator 7805	276-1770	1
CR1	LED, flashing	276-036	1
CR2,CR3	LED, mini, red	276-026	2
CR4	Rectifier, diode, 1N4001	276-1101	1
SW1, SW2, SW3	Switch, submin slide, DPDT	275-407	3
R1,R2	Resistor, 2.2 k, 1/4 watt		2
R3	Resistor, 560 ohm, 1/4 watt		1
R11	Potentiometer, 10-k, 2-watt AF taper	271-1721	1
R13,R14	Resistor, 10 ohm wire wound, 5 watts		2
F1	Fuse holder, 5×20 mm fuses	270-362	1
J1	Standard 4-pin mic socket	274-002	1
J2	3-wire stereo jack, open	274-279	1
J3	Polarized, 3-pin power socket		1
J4,J6	Submin 2-wire jacks	274-292	2
J5	Min 2-wire jack, closed	274-296	1
J7	DC power jack, coaxial	274-1563	1
	Insulated mounting kit for U4	276-1373	1
U2	Optional 8-watt audio amplifier module, GC Electronics, catalog J4-596.		
	Case, aluminum 3″×3″×4″		1

Set of 3 pc boards etched and drilled may be obtained from:
R.& R. Associates, 3106 Glendon, Los Angeles, California 90034, $6.75 ppd.

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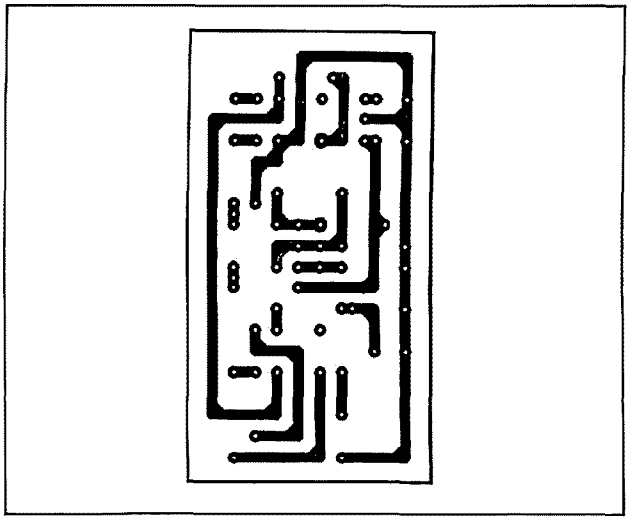
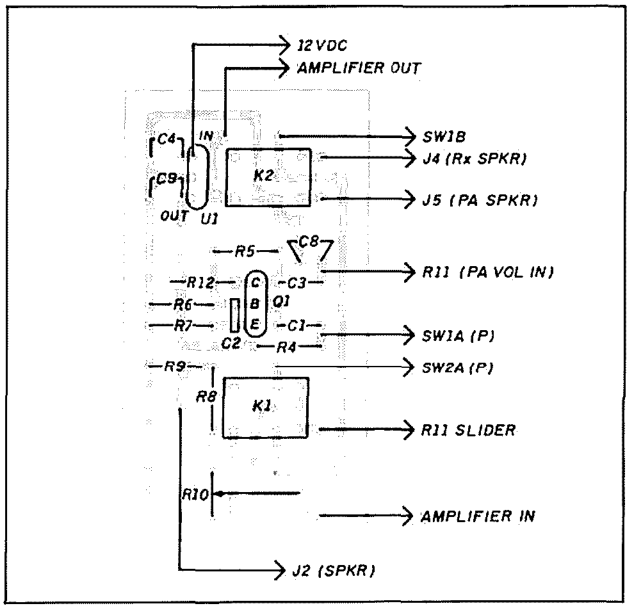
FIGURE 2A



PC board 1, controller, etch side.

FIGURE 2B

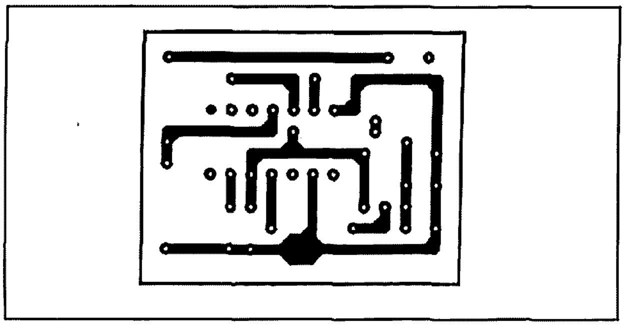


PC board 1, controller, component layout.

Construction

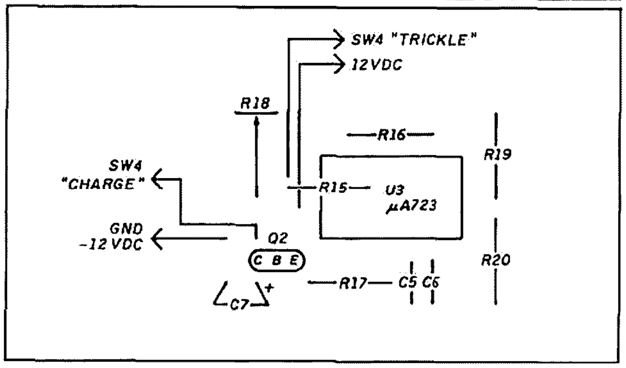
Most of the circuitry is on two boards. You'll need a third board for the 2-watt audio amplifier if you use it instead of a commercial amplifier module. Board 1 includes the mic preamp, K1 and K2 relays, receiver audio-gain control R9 (preset trimmer), and 5-volt regulator U1. Board 2 includes all the parts for the battery charger. Both boards can be assembled on perfboard using point-to-point wiring. For those who prefer them, pc board layouts are shown in Figures 2, 3 and 4. R10 (PA volume), SW1, SW2, SW3, and J1 and J2 are panel mounted.

FIGURE 3A



PC board 2, charger, etch side.

FIGURE 3B



PC board 2, charger, component layout.

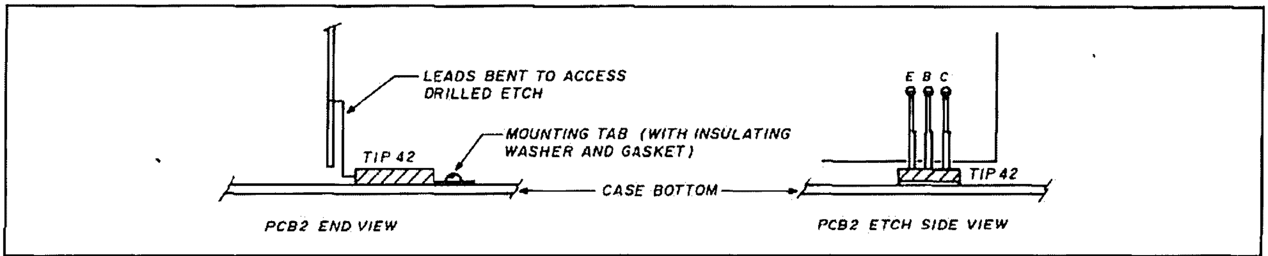
ground bus; all other signal grounds must be connected to it to avoid instability.

Components located on the boards are shown within the dotted lines on the schematic, **Figure 1**. If you want to use pc boards, you can make them using the pc board drawings or obtain them from the source shown in the parts list. If you decide to use point-to-point wiring, these drawings will help you with component layout. Board 1 (controller) is shown in **Figure 2A** (etch side); **Figure 2B** shows the component layout and interconnecting wires. **Figures 3A, B, and C** show board 2 (charger), *and* **Figures 4A, B, and C** the 2-watt audio amplifier. For convenience, the parts list is subdivided by board number or chassis/panel mounting.

Assemble all resistors and capacitors for board 1; then solder them and clip excess leads. *Note that R10 (PA volume) shown in the* **Figure 1** *schematic is not mounted on the board, but on the front panel.* It's connected to the board with miniature shielded cable. Connect the shields to the board ground at one end and to the R10 ground lug at the other. Finally, assemble active components Q1 and U1, solder, and clip excess leads.

Follow the same steps for assembling and soldering board 2. However, note that TIP42 power transistor Q2 is connected to the board on the etch side. The leads are bent so that the device can be mounted on the case bottom for heat sinking. **Figure 3C** illustrates this assembly. Q2 is mounted to the case with a mica insulating washer and no. 4 screw shoulder washer, so that the collector tab is insulated from the case. The mounting hardware is available as a kit, shown in the parts list. After you've put everything together, check your work with an ohmmeter to make sure that Q1 tab is insulated from the case. If it isn't, the Q2 output will be shorted to ground and may damage the device.

FIGURE 3C



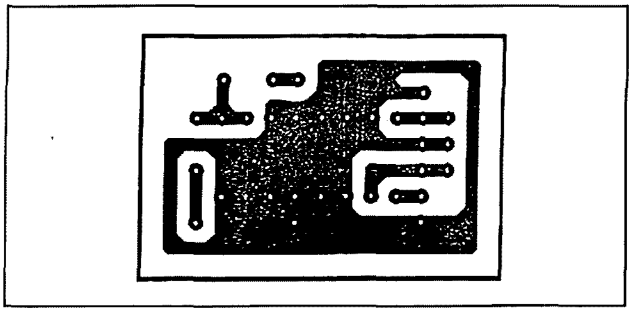
Mounting TIP42 transistor to pc board 2.

When assembling the boards, allow adequate lengths of input/output wire (no. 22 insulated) to reach the destination point on another board or panel control. Use miniature shielded wire (where shown) for connections to the amplifier input and output. It's important to avoid ground loops, particularly with the high gain associated with the preamp and 8-watt amplifier. **Figure 5** shows a "star-type" grounding hookup. This hookup avoids daisy-chain ground loops, which can cause oscillation. The incoming −12 volts DC power line connects to a ground lug mounted to the case with external tooth washers and hardware for low-resistance connections. A no. 20 stranded, insulated wire is run from this lug to the V− input on the amplifier board or module. This is the main

LEDs CR1, CR2, and CR3 are mounted on the front panel. Mounting-hole size depends upon the LEDs used. The mounting holes should allow the LEDs to fit snugly when inserted from the rear of the panel. Once the diodes are installed, put a drop of Crazy Glue™ on the rear panel and also on the LED. (This anchors the LED to the rear panel.) Connect one lead of R1, R2, and R3 to the respective LED *anode* lead. Do this by twisting the leads, soldering them, and snipping off the excess. Connect the LED cathode leads in common with short lengths of no. 22 wire, and then to the case ground. Connect the other LED leads directly to SW1, SW2, and SW3, as shown in **Figure 1**.

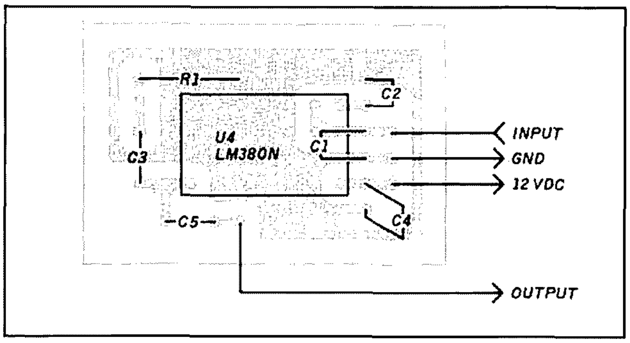
Figure 6A shows a hole-drilling guide. The case I used is 3″ × 3″ × 4″ aluminum. Radio Shack no longers stocks this size, but catalogs the same type in size 3″

FIGURE 4A



PC board 3, 2-watt amplifier, etch side.

FIGURE 4B



PC board 3, 2-watt amplifier, component layout.

(Figure 6A) if you exclude the PA feature. Apply the panel markings after drilling all the holes; use a dry transfer kit. Don't try to do this after mounting the components to the panels. It's a good idea to let the lettering dry overnight before putting on any kind of protective coating. If you don't, the liquid coating may cause the lettering to float off the panel surface. You can use Datakoat™ or clear nail varnish to cover the lettering; it should be *flowed on* gently with a small brush. I prefer two light spray coats of Varathene™ plastic lacquer applied evenly to the whole panel. This dries to a glass-hard finish, which resists scratching.

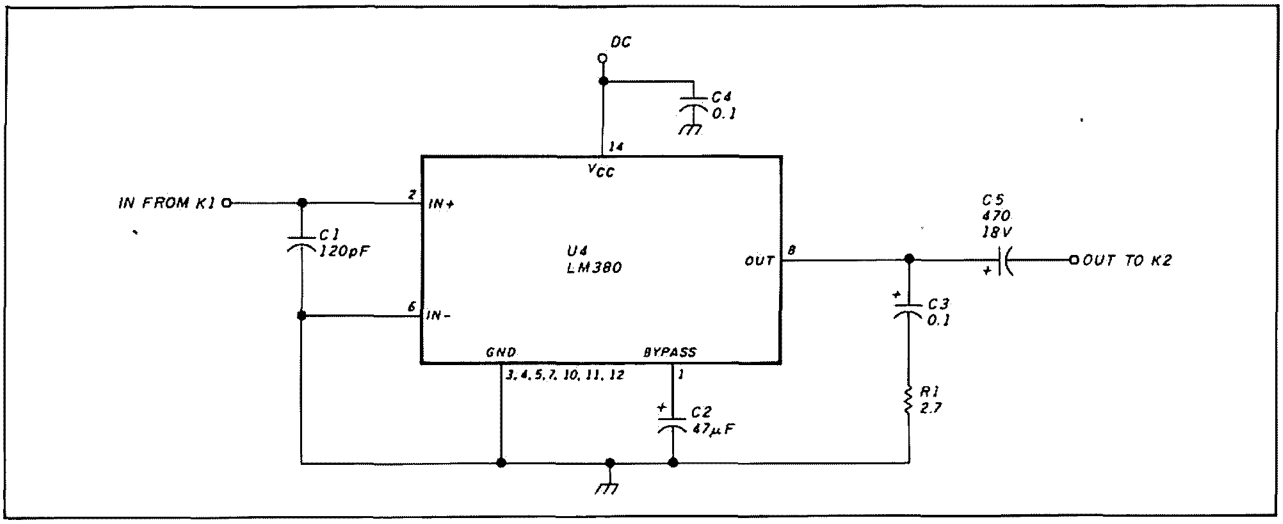
Cable assembly

Your final step is to make the cable which connects the Mobile Companion to your HT. **Figure 7** shows this cable. It's made of two lengths of miniature single-conductor shielded wire and one length of two-conductor shielded wire.

The connectors lor the Companion end are determined by the jacks mounted on the Companion, as shown in the drawing. Select the connectors for the other end based on your radio's connectors. Some HTs have coaxial jacks for external power input with the center pin grounded and V+ on the outer ring. Others have V+ on the center pin and the outer ring is grounded. The Companion power jack has the center pin at V+. If your HT input is center-pin grounded, you'll have to be careful when making a polarity reversal in this cable. Connect the shield to the center pin and wire to the sleeve, as shown in **Figure 7**.

Cable assembly length is determined by the distance between the Companion and the HT inside the car. You can bundle the finished cables into a single assembly with plas-

FIGURE 4C



Schematic of 2-watt amplifier.

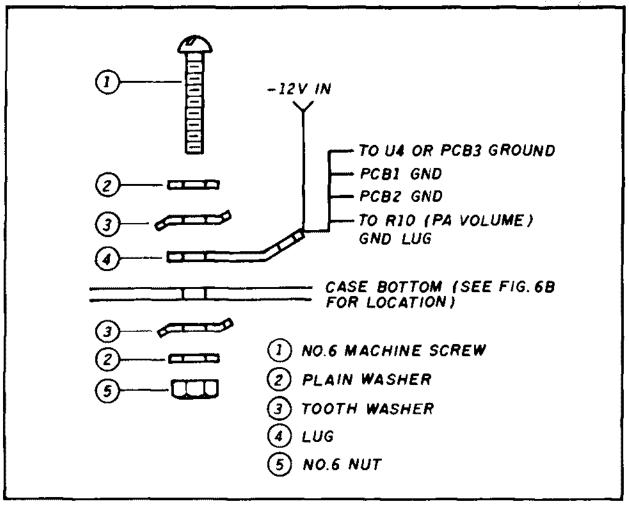
× 5-1/4" × 5-7/8". If you use this case you must modify the drilling guide. I like the latter because it allows more space for ease of construction. Each board is mounted with a single small L-bracket to the case bottom (**Figure 6B**). The 8-watt audio amplifier (if used) is mounted with two holes and 1/4-inch standoffs. You won't need the holes marked with an X

tic cable ties, dressing the length of each to reach the jacks on the HT.

Checkout and adjustment

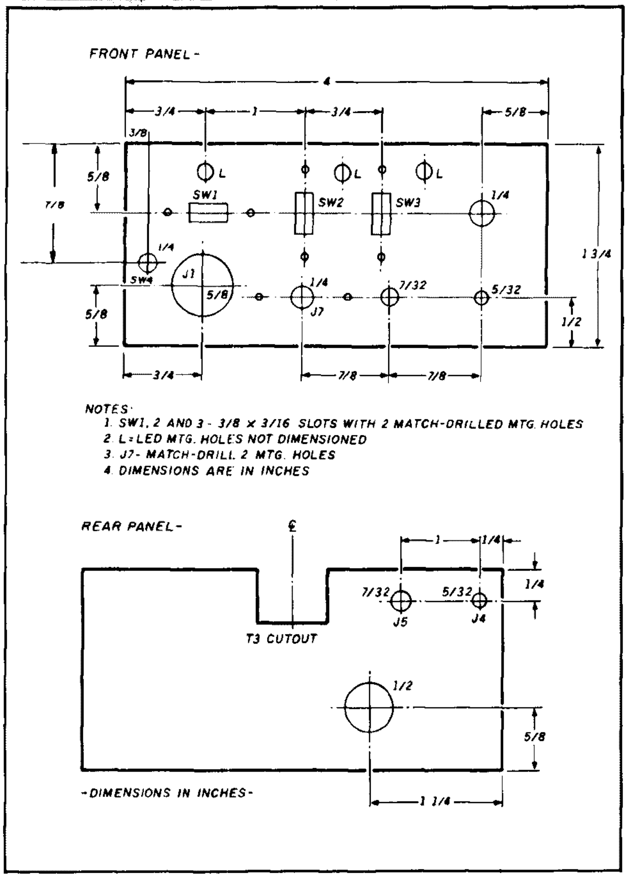
I use a small 0 to 20-volt 200-mA power supply for my initial checkout of any project. This ensures that any shorts or

FIGURE 5



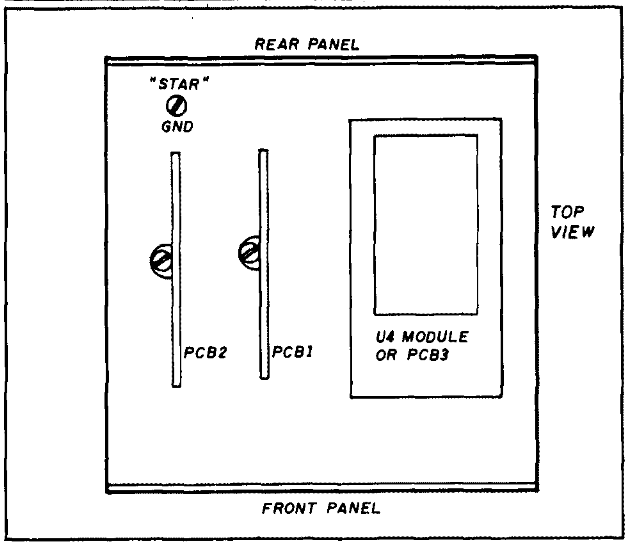
"Star ground" arrangement.

FIGURE 6A



Case drilling guide.

FIGURE 6B



Case component locations.

miswirings have minimal chance of causing damage. Apply 13.6 volts to the input terminals. Check that the U1 output is 5 volts DC and the U4 output is 10.5 volts. If they check out, it's safe to assume that there are no power shorts.

Adjusting HT power

If the voltage measured at U4 OUT is more or less than 10.6 volts DC, vary the value of R21 until the voltage is 10.6 volts DC. If your HT is normally 12 volts DC, you can change R21 to obtain 12 volts regulated output. However, to stay in regulation, the input from the car battery will have to be at least 14 volts (2 volts more than the regulator output). When the engine is idling or stopped, the battery voltage falls to about 12 volts. If your HT has a 6-cell battery pack (7.5 volts), adjust R21 to about 120 ohms to measure 7.5 volts from U4.

Adjusting the battery charger

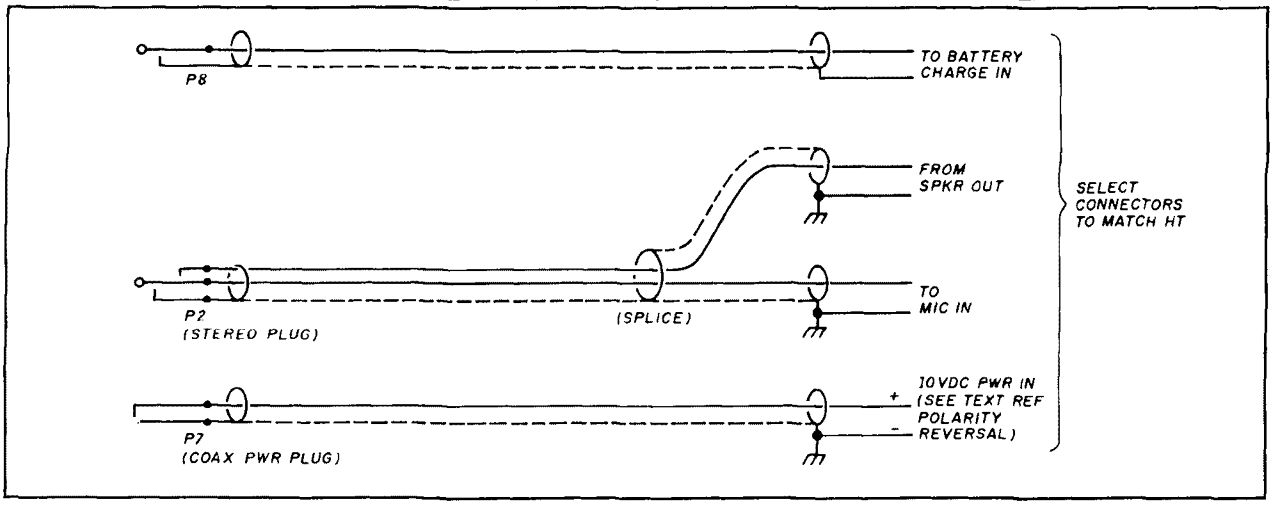
Note that CR4, a 1-A diode, is connected in series with the charger output and J8. CR4 is wired directly from SW3A to the connector. The purpose of this diode is to prevent the battery from shorting when the line is plugged into the HT. If your HT has an internal diode for the same reason, omit CR4 and run a wire between SW3A and J8.

Measure the voltage at SW3A (Q2 collector) with a digital voltmeter, making sure there is no battery connected to J8 and that SW3A is switched to CHARGE. Adjust R18 until the voltmeter reads 12 volts. This is equal to eight cells fully charged to 1.43 volts, plus the 0.7-volt drop through CR4. If you don't use CR4, adjust for a DVM reading of 11.4 volts.

If your HT has a six-cell, 7.5-volt battery, adjust R18 for a DVM reading of 9.3 volts with CR4 in the circuit, or 8.6 volts if CR4 is not used.

Once the unit is set up for your battery voltage, you don't need to make any further adjustment of R18. Charge taper is automatic. You can check current limiting by connecting a 10-ohm, 10-watt wire-wound resistor across J8 in series with a 1-A current meter. You should get a reading of 650 to 850 mA. For a final check, charge the battery fully with its regular charger.

FIGURE 7



Interconnecting cable assembly.

Connect it, in series with a 100-mA current meter, to J8. It should read approximately 25 mA. Switch SW3A to TRICKLE; this should give a current reading of 10 mA. If the reading is higher, increase the value of R15 (100 ohms) to get a reading of 10 mA or less.

One last note: If your HT doesn't have separate inputs for POWER and CHARGE, you obviously can't use the setup as described. In this case you may omit the U4 regulated power circuit and cable, but have the charger connected during operation. The HT will operate from its battery during transmit mode, and the battery will receive a "boost" charge during receive cycles. This is an acceptable alternative that I used for several years without having a battery run down on me.

Adjusting the receiver audio amplifier level

R10, the trimmer potentiometer on pc board 1, is a set-and-forget control for adjusting the audio level to the external speaker. Connect the HT speaker output to J4. Make sure SW1 is set to RADIO. Adjust R10 fully counterclockwise. Turn on both units. Set the HT to an active channel. Turn the HT volume fully clockwise. Adjust R10 until audio level from the external speaker is a little too high for comfort. A comfortable audio level is always regulated by the HT volume control; R10 needs no further adjustment.

Checking the PA system (optional)

Plug the microphone into J1 and PA, or a test speaker into J5. Set SW1 and SW2 to PA. Set R11 (PA VOLUME) counterclockwise. Turn on Companion. Advance PA volume slowly clockwise while talking into a microphone. **Caution:** If a test speaker is close by, you will get loud audio feedback. Run the speaker in another room, if possible. Disconnect the mic. Advance PA volume fully clockwise; an increasing hiss should be heard.

If an audible howl occurs near full volume, or if the hiss level drops suddenly near full volume-control setting, instability is present. The drop in the hiss level is caused by high-frequency oscillation above the audible level. If this happens, check U2 output with an oscilloscope or AC voltmeter for obvious oscillation. If it is present, it's almost certainly due to poor grounding. Check to see that your grounding conforms with that shown in **Figure 5**. *Do not let the 8-watt amplifier continue to run in an oscillating mode. Severe overheating and damage will occur.*

BIBLIOGRAPHY

1. Joe Moell, KØOV, "A Fast Charger and Regulator for the Tempo S," QST, June 1981, page 35

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WANTED: Ham equipment and other property. The Radio Club of Junior High School 22 NYC, Inc. is a nonprofit organization, granted 501(C) (3) status by the IRS, incorporated with the goal of using the theme of Ham Radio to further and enhance the education of young people nationwide. Your property donation or financial support would be greatly appreciated and acknowledged with a receipt for your tax deductible contribution. If you call or write today it's not too early to get a 1990 tax deduction. It's easier, faster, and usually more profitable to donate than to sell. Most important, you're helping. Write us at: PO Box 1052, New York, NY 10002. Round the clock hotline: (516) 674-4072.

JOHNSON VIKING RANGER Transmitter/exciter, and Drake 2-B receiver—excellent—$200.00. Stephen Hollis, KA9SSP, 1876 Adams Rd, Eagle River, WI 54521. (715) 479-4604.

FOR SALE: One Delta Ten-Tec Transceiver with Delta antenna tuner, mic and power supply. $400.00. H. Nelsen (616) 642-6573.

FREE LIST of low cost ham equipment. Write to Jim Braddy, WA4DSO, 3037 Audrey Drive, Gastonia, NC 28054.

WANTED: All types of Electron Tubes. Call toll free 1-800-421-9397 or 1-612-429-9397. C & N Electronics, Harold Bramstedt, 6104 Egg Lake Road, Hugo, MN 55038.

TEKTRONIX: 7603N11S, Two 7A15AN11, One 7B53AN11 plug-ins (no digital readout), panel cover. No probes. $450. ($425 pickup). General Dynamics: R1051B/URR receiver, 2-30 MHz $500. Collins: 51S1 $850. 516E-1 $50. James Craig, W1FBG (603) 964-6658.

DIGITAL AUTOMATIC DISPLAYS. Any radio. Be specific. Large 45 cent SASE. GRAND SYSTEMS, POB 2171, Blaine, WA 98230.

RADIO SHACK Color Computer Ham Software and Hardware. Free catalog. Dynamic Electronics, POB 896, Hartselle, AL 35640. (205) 773-2758.

WANT OLD BUGS (Speed keys) of any type. WB4EDB, Smiley While, PO Box 5150, Fredericksburg, VA 22403.

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POLICE/FIREFIGHTER HAMS. Please send your call, name, address, rank, department name for inclusion in special roster, available early 1990. Capt. Bob Blakeslee, N2IHQ, 1-1/2 Macomber Avenue, Binghamton, NY 13901.

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AVANTEK ATF10135, $12.00, MMIC's, P.C. board, SASE: WA3IAC, 7148 Montague St, Philadelphia, PA 19135.

"HAMLOG" COMPUTER PROGRAM. Full features, 17 modules. Auto-logs, 7-band WAS/DXCC. Apple $19.95. IBM, CP/M, KAYPRO, Tandy, C128 $24.95. HR-KA1AWH, POB 2015, Peabody, MA 01960.

YAGI BUILDERS. 6061-T6 tube traps. Good for 1500 PEP. SASE for details. No collect calls. BROWN ENGINEERING INC, 5501 SW 25th Court, Hollywood, FL 33023. (305) 989-4658.

KENWOOD TR-9000 2m all mode, factory boxed $325. Mobile 8 channel receiver 30-50 MHz $10. VHF mobile repeater/extender unit w/manual $75. Several old Hallicrafters receivers $25—$75. WANT: Yaesu FT-2700RH mobile, Cushman CE-3 test set. K6KZT (805) 528-3181.

HAM SWAPPER TRADING POST. FREE first ad up to 30 words. Purchase not required. Additional words 20 cents each. Twice monthly publication of Ham and Computer related equipment. Subscriptions $11.00/year (24 issues). Ham Swapper, 217 N. Hamilton, Pratt, KS 67124.

WANTED: Schematic or any other information concerning the AN/URR58 receiver from Squires-Sanders. Wanted also SS-1R, SS-1VBS in good condition;. Marion Arthur Ambrose Jr, Contel Page Inc, APO New York, NY 09794-007.

450 MHz SPECTRUM ANALYZER. Adapted from Nov 85 QST article by Al Helfrick, K2BLA. Features: Three digit LED center frequency digital readout. 12 position calibrated Scan Width, 1 kHz to 50 MHz, switchable bandwidth; Wide ⁵ 300 kHz, Narrow ⁵ 10 kHz. Use your low frequency scope for the display portion. Switchable 10 kHz video filter. Variable IF attenuator, LOG output calibrated in 10 db steps. For complete kit, order #450-KIT $459.95 plus $4.50 s/h. For assembled and tested unit, order 450-AST, $799.95 plus $6.50 s/h. Calif residents add 6% sales tax. Foreign orders add 15% for shipping. A & A Engineering, 2521 W. LaPalma, #K, Anaheim, CA 92801. (714) 952-2114.

IMRA International Mission Radio Association helps missionaries. Equipment loaned. Weekday net, 14.280 MHz, 1-3 PM Eastern. Nine hundred Amateurs in 40 countries. Rev. Thomas Sable, S.J., University of Scranton, Scranton, PA 18510.

FOR SALE: Kenwood R-2000 SW receiver, $400. Ranger AR-3500 10 meter, 100W mobile XCVR, $300. Both excellent condition and appearance. Matt, WA1HRE, (203) 693-0468/693-6596. Leave message.

INTERESTED IN PUBLIC SERVICE? Join your Local Radio Emergency Associated Communications Team. In Pennsylvania call (717) 938-6943 or write REACT, 1160 Old Trail Rd, Etters, PA 17319.

PC DRILL BITS sizes 55, 56, 58, 60, 62, 65, 66, 68, 70. $1.25 each or 5 for $5. Free shipping. S Systems, Box 1724, Tempe, AZ 85280-1724.

ICOM IC-04AT UHF Handie Talkie. $260.00 postpaid. Postal money order only. Rick Blackburn, 433 So. Wall #4, Los Angeles, CA 90013-1516.

RUBBER STAMPS: 3 lines $5.00 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

ELECTRON TUBES: Receiving, transmitting, microwave... all types available. Large stock. Next day delivery, most cases. DAILY ELECTRONICS, PO Box 5029, Compton, CA 90224. (213) 774-1255.

RECONDITIONED TEST EQUIPMENT $1.25 for catalog. Walter, 2697 Nickel, San Pablo, CA 94806.

COMING EVENTS

Activities — "Places to go . . ."

SPECIAL REQUEST TO ALL AMATEUR RADIO PUBLICITY COORDINATORS: PLEASE INDICATE IN YOUR ANNOUNCEMENTS WHETHER OR NOT YOUR HAMFEST LOCATION, CLASSES, EXAMS, MEETINGS, FLEA MARKETS, ETC, ARE WHEELCHAIR ACCESSIBLE. THIS INFORMATION WOULD BE GREATLY APPRECIATED BY OUR BROTHER/SISTER HAMS WITH LIMITED PHYSICAL ABILITY.

ONTARIO: February 3. The Niagara Peninsula ARC's 12th annual Big Event, hamfest and dinner dance, C.A.W. Hall, 124 Bunting Road, St. Catharines. Admission $3. Talk in on 147.24/84. For information write NPARC, PO Box 692, St. Catharines, Ont. L2R 6Y3 or phone (416) 682-4844.

NEW YORK: February 11. Melville. Hamfest sponsored by the Long Island Mobile ARC, Electricians Hall, 41 Pinelawn Rd. 9 AM to 4 PM. $5 at door. No advance. Talk in on 146.25/85. Contact Neil Hartman, WE2V (516) 462-5549 or Mark Nadel, NK2T (516) 796-2366.

OHIO: February 11. The Mansfield Mid*Winter Hamfest/Computer Show, Richland County Fairgrounds, Mansfield. Doors open 7 AM. Tickets $4/advance; $5/door. Tables $7/advance; $10/door. Talk in on 146.34/94. For information of tickets/tables SASE to Dean Wrasse, KB8MG, 1094 Beal Road, Mansfield, Ohio 44905. (419) 589-2415 after 4 PM EST.

MASSACHUSETTS: February 17. Electronic Flea Market sponsored by the Algonquin ARC. Marlboro Middle School Cafeteria, Union St, Marlboro. 10 AM to 3 PM. Sellers 8 AM. Admission $2.00. Tables $10/advance; $12/door. Wheelchair accessible. For information Ann, KA1PON (508) 481-4988 or write AARC, Box 258, Marlboro, MA 01752.

FLORIDA: February 17-18. Sarasota Hamfest. South Florida Section ARRL Convention, Computer Show, Robarts Arena, 3000 Rigling Blvd, Sarasota. 9 AM to 4 PM. Admission $5/advance; $7/door. Talk in on 146.31/-.91, 147.90/-.30 and 449.425/-.925. For information contact Hadley Carrigan, N4ODK, 101 N. Adams Drive, Sarasota, FL (813) 388-2868.

INDIANA: Saturday, February 24. The LaPorte ARC's Winter Hamfest, LaPorte Civic Auditorium. Donation $3.50. Talk in on 146.52 simplex. For table reservation ($3.50) SASE to LPARC, POB 30, LaPorte, IN 46350.

MINNESOTA: February 24. The 9th annual Midwinter Madness Ham Radio and Computer Show, Medina Ballroom, Highway 55. Sponsored by the Robbinsdale ARC. 7 AM to 2 PM. Admission $3/advance; $5/door. Talk in on 147.60/00. For information Bob Arel (612) 545-1059 evenings and (612) 425-2424 days.

FLORIDA: February 24. Brooksville. The Hernando County ARA's 8th annual Hamfest, Hernando County Fairgrounds auditorium. Tickets $3/advance; $4/door. Swap tables $8. For tickets/table reservations SASE with check to Hernando *County Amateur Radio Association,* PO Box 34605-1721, Brooksville, FL 34605.

VERMONT: February 24. Milton. The Northern Vermont Mid-Winter Hamfest, Milton High School, Rt 7. 9 AM to 3 PM. Admission $2. Free tables FCFS. Talk in on 145.47/-600. For information Mitch Stern, WB2JSJ (802) 879-6589 or Tom Taylor, N1EXY (802) 893-4834.

TEXAS: February 24. The Orange ARC's fifth annual Hamfest-Flea Market, National Guard Armory, Meeks Drive, Orange. 8 AM to 4 PM. For information Sherwood Buckalew, KA5VOT (409) 883-6111, Dan Killough, WB4GYS (409) 769-9603 or Rick Lewis, KF5GV.

MICHIGAN: February 25. The 20th annual Livonia ARC's Swap 'n Shop, Dearborn Civic Center, Dearborn. 8 AM to 4 PM. Talk in on LARC Repeater 144.75/5.35 and 146.52 simplex. For information SASE to Neil Coffin, WA8GWL, Livonia ARC, POB 2111, Livonia, MI 48151.

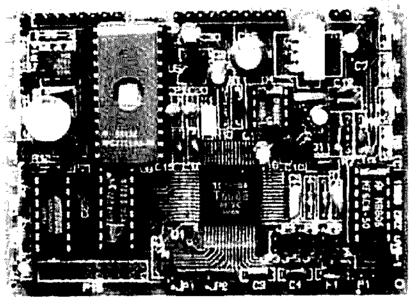
VIRGINIA: February 25. Winterfest sponsored by the Vienna Wireless Society, Vienna Community Center, 120 Cherry Street, Vienna. Setup 6 AM. Public 7:30 AM. Admission $5/door. $10/tailgate. No advance. For information Harry Kaklikian, W4ACN, 4941 Andrea Avenue, Annandale, VA 22003. (703) 978-4402.

OHIO: February 25. The Cuyahoga Falls ARC's 36th annual Hamfest, Akron North High School. 8 AM to 3 PM. Tickets $3/advance; $4/door. Tables $5/advance; $6/door. Wheelchair accessible. For information Bill Sovinsky, K8JSL, 2305 24th St, Cuyahoga Falls, OH 44223. (216) 923-3830.

(continued from page 71)

Model SVB-16

The SVB-16 features 16 soft-sectored message cells and can store up to 100 seconds of digital voice audio. The unit has a 40-kHz digitizing rate combined with 14 poles of audio filtering. A CW memory keyer option is available. This option gives you a combination 16-message voice/CW memory keyer. Each has separate soft-sector digital memories and automatic serial number insertion. CW messages are stored in a nonvolatile EEPROM, and voice messages are retained by an optional battery when power is disconnected. If you have a personal computer, you can store and retreive voice and CW memory templates on your PC's hard disk via a high speed serial port.



For more information on any of these units contact QRZ Industries, PO Box 160, Piedmont, South Carolina 29673. Phone: (803)269-0000.

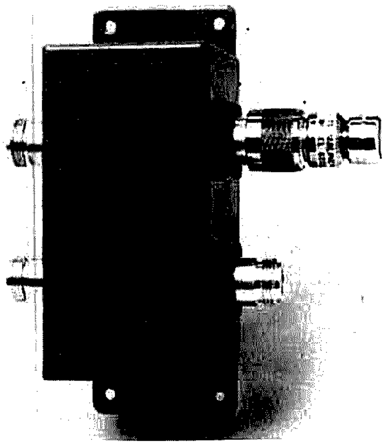
Circle #305 on Reader Service Card.

Unique, High Power Bidirectional Coupler Available from Bird

Bird Electronics Corporation offers the Model 4266 high power, bidirectional coupler with a continuous rating of 1500 watts CW in the primary line. The coupler is useful for power measurement, frequency measurement, automatic level control, automatic frequency control, VSWR measurement, attenuation measurement, and signal observation.

With a nominal coupling flatness of ± 0.25 dB over the entire frequency range (typically ± 0.1 dB), the 4266 provides a signal sample of forward and/or reflected power decoupled by 30 dB from the mainline. It has 0.1-dB insertion loss and VSWR below 1.05. The 4266 allows reversed input and output and provides simultaneous availability of forward and reflected signal samples from the secondary line. For complete details, contact Bird Electronic Corporation, 30303 Aurora Road, Solon, Ohio 44139. Phone: (216)248-1200.

Circle #306 on Reader Service Card.



IC-4KL Solid State Linear Amplifier

ICOM's new IC-4KL linear amplifier is now available. The 1000-watt full function IC-4KL features:

- 1000 watts output at full duty cycle
- Separate controller and amplifier
- Fully automatic, high speed antenna tuner
- Full break-in operation
- Complete circuit protection

Additional features include: built-in 40-volt 60-A power supply with automatic selection for use with 100 to 120 volt AC or 220 to 240 volt AC power, two needle meters that register SWR, a final transistor current, output power, and final transistor temperature. Also included are rack mounting handles for mounting with a 19-inch rack. The IC-4KL connects quickly to any ICOM transceiver! The suggested retail price of the IC-4KL is $6995.

For information contact ICOM America Inc., PO Box C-90029, 2380 116th Avenue NE, Bellevue, Washington 98009-9029. Telephone (206)454-8155 or FAX (206)454-1509.

Circle #307 on Reader Service Card.

ICOM IC-RP1510 144-MHz Repeater

ICOM has introduced the UC-RP1510 repeater with local or remote control, automatic battery reverting, and extensive RF shielding

The IC-RP1510 features:

- 25 watts output power
- Broad band frequency coverage
- DTMF remote control
- Dip switch selectable squelch
- Built-in external battery connection
- Extensive RF shielding

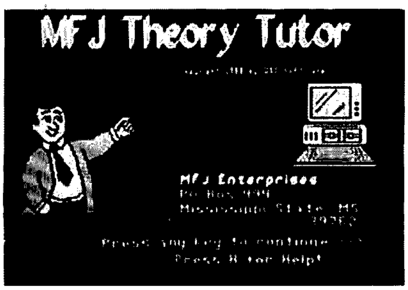
The IC-RP1510 repeater has an adjustable time out timer, a programmable callsign identifier, and a built-in speaker and microphone connecter. The suggested retail price is $1,849.

For details contact ICOM America, Inc., PO Box C-90029, 2380 116th Avenue NE, Bellevue, Washington 98009-9029. Telephone (206)454-8155, FAX (206)454-1509.

Circle #308 on Reader Service Card.

Theory Tutor for IBM Compatibles

MFJ's new Theory Tutor for IBM compatibles helps you study for the theory portion of any FCC license exam. Study the entire FCC question pool, concentrate on selected areas, or try taking sample tests. Each study session is saved automatically, and you can return to a previous session at any time. Print out tests on any Epson or IBM compatible printer. The program includes graphics with appropriate questions, a complete scoring analysis, color change options, an on-line calculator, explanations of hard questions, and more.



The MFJ Theory Tutor is available for each license class. Ask for model numbers: MFJ-1610-Novice, MFJ-1611-Technician, MFJ-1612-General, MFJ-1613-Advanced, or MFJ-1614-Extra at $29.95 each. For more information contact MFJ Enterprises, Inc., PO Box 494, Mississippi State, Mississippi 39762 or call (601)323-5869. FAX: (601)323-5869.

Circle #309 on Reader Service Card.

Elmer's Notebook

Tom McMullen, W1SL

ELEMENTARY ELECTRONICS: CAPACITORS

In previous columns I've discussed magnetic fields and resistance. This month I'd like to explore another important element in electronics — capacitors and how they work.

In simple terms, a capacitor is an energy storage device; it stores electrons. The number of electrons a capacitor can store, and how long they remain stored, depends on such things as the size of the electrodes (more surface area allows more electrons to be stored) and the type of dielectric (material between the electrodes) used. For long term storage, the dielectric must not act like a resistor. This means it can't let the electrons "leak" from one plate to the other.

What's in a capacitor?

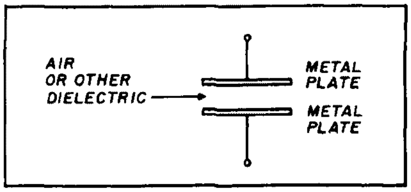
There are several ways to make a capacitor. If you've ever looked through one of the "grab bag" assortments at a supply house or hamfest flea market, you know that capacitors come in a variety of shapes and sizes. A very basic capacitor is simply two metal plates or disks parallel to each other, with air between (see **Figure 1**). In this type of capacitor either plate can be of either polarity (+ or −), and the air between is called the dielectric. The dielectric material between the plates acts as insulation to keep the plates apart and concentrates the electrostatic lines of force between them. The standard dielectric constant for air is approximately 1.006 (vacuum is rated at 1.0). Other materials have higher numbers. This means that if you have a pair of plates with air between and a capacitance of 100 pF, and you replace the air with some material that has a dielectric constant of 2.0, you end up with a capacitance of approximately 200 pF for the same size. The capacitance is directly proportional to the area of the plates and the dielectric constant of the material between them. It increases when you decrease the spacing between the electrodes or plates. Paper, glass, ceramic, and many plastics are types of dielectric materials used in capacitors. Each material has characteristics making it desirable for one use or another. **Table 1** gives the constants of some of the most common materials.

Some capacitors, called electrolytics, have a paper or plastic material between the electrodes that is impregnated with a chemical which increases the dielectric constant to many times that of air. This allows a much greater "storage capacity" than is found in a capacitor of the same size using plain paper, glass, ceramic, or air dielectric. Such units are usually polarized, and any voltage applied must be of the correct polarity to prevent damage to the capacitor or power supply. If the applied voltage has the wrong polarity, the dielectric will usually overheat and short circuit. When too much voltage or reverse voltage was applied to the older "wet" electrolytic capacitors, they would "cook" and release a jet of steam with a startling bang. Modern electrolytic capacitors use a dry electrolytic which usually isn't prone to this type of destruction, but simply becomes a short circuit.

What's a pF?

The basic unit of capacitance is a farad, named after physicist Michael

FIGURE 1



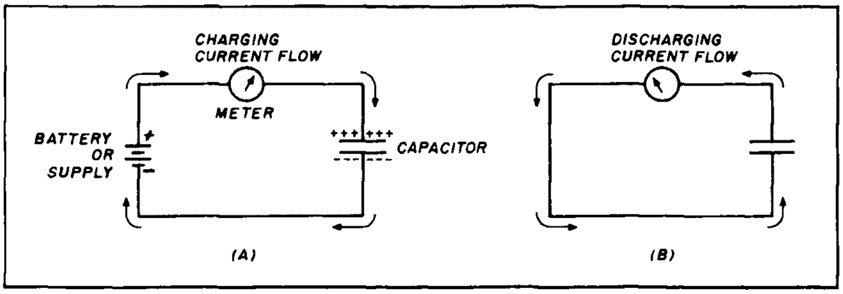
A simple representation of a capacitor — two metal plates with air or some other insulation (dielectric) between them.

TABLE 1

Dielectric constants for some common materials.

Vacuum	1.000
Air (dry)	1.006
Rubber	2 or greater
Paper	2 or greater
Ceramic	3 to 7
Glass	4 to 8
Mica	6 or 7
Water (pure)	80
Barium Titanate	7,000 or higher

FIGURE 2



When a capacitor is connected to a power source, electrons flow from one plate into the other, causing an instantaneous charging current (A). When the capacitor discharges through an external circuit, the current flow reverses (B).

Faraday. A farad is a very large unit. The "official" description says that a capacitor "has a capacitance of 1 farad when a change of 1 volt per second across the unit produces a current of 1 ampere through it." Another definition says that a 1-F capacitor can store a *difference* of 1 coulomb (6.25 $\times 10^{18}$ electrons) between the plates. These definitions tell us something about "changing" voltages and current, but what does this mean in practical terms? In the days before modern materials were developed for miniaturization, the saying was that a 1-F capacitor (they were cometimes called "condensers") would be about the size of a one-car garage. More recently, someone estimated that a medium-sized 12-volt automotive battery had the approximate storage capacity of a 1-F capacitor. A farad is also far larger than anything we work with in electronics, so rather than write out all the decimal fractions of a farad for our circuits and calculations, we use abbreviations. One common unit is a microfarad, abbreviated μF. Another is a picofarad, or pF. There is also a nanofarad, nF, which is not widely used in the United States.

Here's what those abbreviations mean in decimal terms:

1 μF = 0.000,001 farad
1 nF = 0.000,000,001 farad
1 pF = 0.000,000,000,001 farad

You can see why the abbreviations are so helpful! As a general rule, the larger units — like 1 μF to several hundred μF — are used to filter out noise, hum, or ripple in power supply circuits and to pass lower audio frequencies from one stage to another. The smaller values, from 1 pF to .01 μF, are often used in tuned circuits or in critical coupling between radio frequency circuits.

How capacitors work

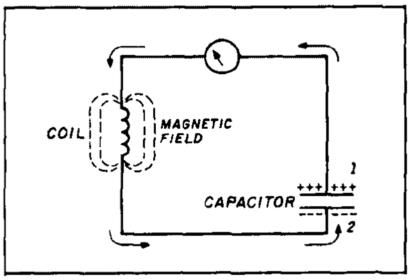
As I noted before, a capacitor is a device that stores electrons. When two electrodes separated by a dielectric are connected to a source of power, there's an immediate rush of free electrons from the power supply to the plates (see **Figure 2A**). This is where the "current flow" part of the definition comes into play. Because the electrons are flowing from the supply to one plate, and from the other plate to the supply, current *appears* to flow through the capacitor. Because of the dielectric, the electrons can't cross from one plate to the other. Instead, they build up on one plate, creating a surplus of electrons like that found on the terminal of the battery. The other plate, which is connected to the battery terminal that has a deficiency of electrons, takes on the same characteristic as that terminal. It too has a deficiency. When a circuit is completed between the plates, the electrons try to get to the other side, causing a brief current flow out of the (+) plate to the (−) one. Interestingly enough, the positive (+) terminal is the one that has a deficiency of electrons while the negative (−) terminal has a surplus of electrons. However, we still say that current flows from positive to negative — even though electrons are going the other way. (Old habits are hard to break.)

This means you have two plates with electrons trying to get across the gap, and in this condition the capacitor is said to be "charged." If the dielectric is of good quality (dry air, for example), the charge will remain for a long time — minutes, or even hours — after the power supply is disconnected. However, many dielectric materials aren't that good; the charge leaks off in a few minutes through the resistance caused by moisture and poor insulation used in manufacturing the capacitor.

Taking advantage of the rush

The tendency of electrons to rush from one plate to the other is put to good use in tuned circuits. When a capacitor is connected across an inductor, as in **Figure 3**, something interesting takes place. For purposes of this discussion, imagine that one plate of the capacitor has been given a charge from a power supply. The supply is then disconnected and the inductor is connected across the capacitor. The extra electrons on one plate try to rush to the other side through the wire in the coil. But as I mentioned in an earlier column about magnetic fields, a field is created whenever current flows in a wire, and if the wire is wound into a coil the field is more intense. In addition, the field from the first turn in the coil cuts across the second turn, and the field from the second turn cuts across the third turn, and so on. It so happens that when the magnetic field cuts across the second turn, it generates a small current flow opposite in direction to the current flow in the first turn. This opposes the current flow from the capacitor and increases the time required for the capacitor to discharge. The capacitor eventually wins out and the electrons do get to the other plate, but not as quickly as they would through a short circuit. When the electrons stop moving, the magnetic field around the coil collapses, releasing the stored energy back into the capacitor. As a result, plate two now has too many electrons. This starts a rush to get to the other side again, and the inductor tries to slow things down as before. This process continues until the resistance of the wire and the energy in the magnetic field have depleted the charge of electrons.

FIGURE 3



Current flow is impeded by a coil (inductance) and its magnetic field. The interaction between the coil and capacitor creates a resonant circuit — a vital part of all radio circuits.

The amount of time that the electrons require to move from one plate to the other is an indication of the resonant frequency of the circuit. For instance, if the size of the capacitor (number of electrons stored) and the effect of the inductor (intensity of the magnetic field) allow the electrons to complete their journey 1,000,000 times in one second, it is resonant at one million cycles per second. Older hands will remember this as 1 megacycle (1,000,000 cps), but modern terminology uses hertz as a term meaning "cycles per second," so this circuit is resonant at 1 megahertz, or 1 MHz.

Obviously, changing the size of the capacitor (number of electrons it can store) or the inductor (larger or smaller magnetic field) will change the cycle time of the circuit, thus changing the resonant frequency.

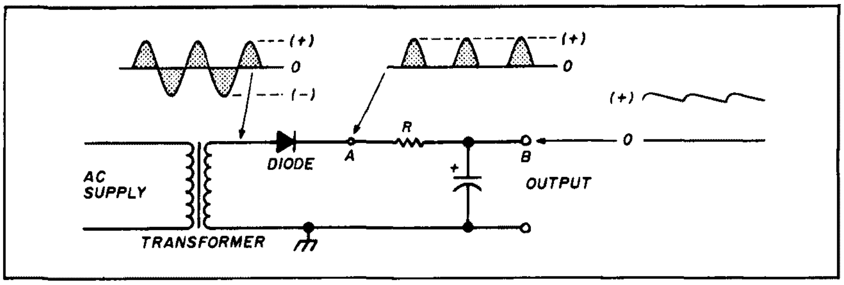
Uses for the larger capacitors

Large value capacitors, especially those of one to several thousand μF,

are widely used in audio and power supply circuits to remove hum and noise and provide the smooth DC required for almost all circuits. The output from the rectifier diode in a simple power supply (point A in **Figure 4**) is simply half of a sine wave. But it's not really useful in that form. An audio amplifier with this type of power applied would have a loud hum. Here's where the capacitor comes to the rescue. In the period when this pulsating voltage is present at the output of the rectifier, the capacitor is storing some of the energy. When the voltage drops to zero between diode conduction cycles, the capacitor releases that stored energy to the output and whatever is hooked to it. The result is a somewhat smoother output, as at point B in **Figure 4**. The larger the capacitor, the more energy it stores and releases, and the smoother the output becomes. It usually requires a very large value to smooth the DC output completely, so some power supplies use two capacitors in a dual section filter (see **Figure 5A**). In this case, the first capacitor does a fair job of smoothing out the "ripples." The current then flows through the resistor to the second capacitor, further smoothing the DC until it reaches a nearly "pure" DC state. In some power supplies, the resistor is replaced by an iron core inductance called a "choke" (see **Figure 5B**). This choke helps by storing and releasing energy in its electromagnetic field. (You haven't heard the last of this magnetic field business. Just wait until I explore how alternating current, or AC, behaves in various circuits in a future issue!)

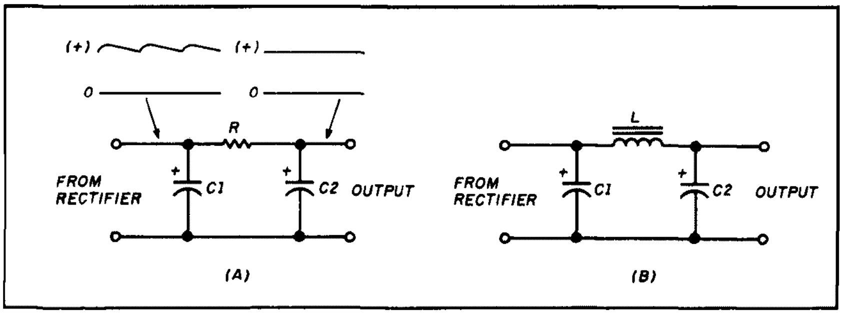
High voltage supplies, like those required by vacuum tube amplifiers, often use a capacitor-choke-capacitor filter to keep the physical size of the required capacitor to a minimum. A 500-μF capacitor that works at 15 volts is often only an inch or so long and a half inch or less in diameter. The same capacitance that would work at 1000 volts or so would be several inches tall and probably several inches wide.

FIGURE 4



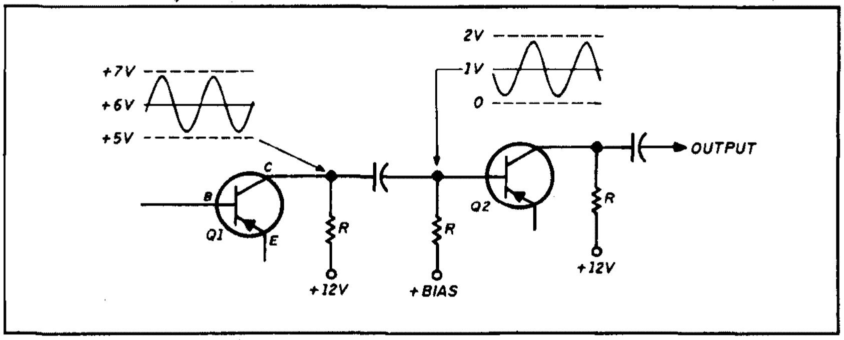
Capacitors also smooth out the rectified AC from a diode in a power supply, making it useful to power other circuitry.

FIGURE 5



Adding a second capacitor and a resistor to the power supply output further smooths the DC output (A). Placing an iron core inductance (filter choke) in the circuit helps smooth the DC output and allows smaller value capacitors to be used (B).

FIGURE 6



Capacitive coupling between two transistor amplifiers. A replica of the audio signal at Q1 is impressed on the bias voltage of Q2 while keeping the two DC voltages separated.

What is a "coupling" capacitor?

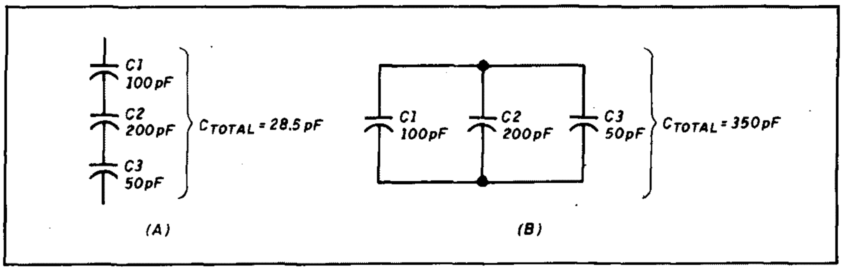
A "coupling" capacitor is obviously used to couple something. If a transformer is used between the first amplifier and the second one in a simple audio circuit, it is said to be "transformer coupled." If a capacitor is used, the circuit is "capacitance coupled" (see **Figure 6**).

When a voltage change appears on one plate, an equal and opposite change appears on the opposite plate. That describes capacitive coupling. Let's follow what happens in **Figure 6** when a signal (audio sine wave) appears at the collector of the first transistor (Q1). As the voltage starts to go positive on the collector, the same thing is happening on one plate of C1 (the capacitor in the center of the drawing). As the electrons increase on one plate, the electrons on the other plate flow back to the power supply through the ground or chassis return. This causes an inverted replica of the signal to appear on the base of Q2. The signals aren't exact opposites of each other because there's a slight lag, called phase shift, through a capacitor. For purposes of discussion, we'll ignore this lag.

When the voltage on C1 (Q1 collector) passes its peak, drops to zero, and

starts going in the opposite direction (negative), the voltage on the base of Q2 follows — producing a replica of the signal on the base of Q2. The transistor then amplifies the signal, and so on. There's an important secondary capacitor function here which you must not overlook. The DC operating voltage on the collector of Q1 is prevented from getting to the base of Q2. This is vital because Q2 needs only a small fraction of a volt on its base to work properly; too much voltage here would destroy the transistor. Because the capacitor passes the *changes* in voltage along, but not the *steady state* DC voltage, the DC bias on the base stays where it belongs and all is well. Thus, you can say that the signal is "coupled" from the first stage (Q1) to the second stage (Q2) by the coupling capacitor (C1). This principle works in the same way for audio, IF, and RF signals. The major difference is that C1 must be larger for audio (typically 0.05 to 5 μF) and smaller for RF (2 pF or less to perhaps 500 pF), depending upon the frequency being coupled. The capacitance required is related to the reactance and impedance in an AC circuit. (I'll discuss this in an upcoming column on alternating current.)

FIGURE 7



Capacitors can be hooked in series or parallel to obtain unusual values. See text for the formulas used.

Series and parallel capacitors

Just as you can use resistors connected in series or parallel to obtain a needed value, you can work with capacitors. However, there is a slight difference. The formulas for capacitance are reversed from those for resistors. Capacitance values in series produce *smaller* values; those in parallel produce *larger* values. **Figure 7** gives some examples.

The formula for series capacitors is:

C(total) = 1/((1/C1)+(1/C2)+(1/C3)),

and so on. Here's a tip. In a series hookup, the total capacitance will always be smaller than the value of the smallest capacitor. Use the formula to calculate how much.

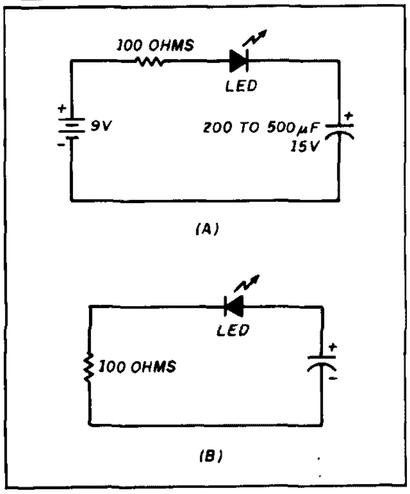
For parallel capacitors, the formula involves straight addition:

C(total) = C1 + C2 + C3, and so on.

The experiment

Here's a simple experiment you can perform to see how a difference in

FIGURE 8



Note the time required to charge the capacitor through the resistor and LED at A, then reverse the diode and note the discharge time. You can try different values of capacitance to see how the time varies for charging and discharging.

capacitance value changes the amount of energy needed to charge a capacitor, and how long it will take to release that energy to an external circuit. The hookup is shown in **Figure 8**. It uses an LED, a resistor (to limit current through the LED to a safe value), and a 9-volt battery. Try some different capacitor values; many grab bags will contain several. Start with a large one — say 400 to 500 μF rated at 15 or 20 volts.

First connect the battery, resistor, and LED together and touch one leg of the LED to the opposite side of the battery. If it lights, you're all set; if it doesn't, reverse the LED wires and try again. After you've proved that the hookup works, connect the negative (–) lead of the capacitor to the negative side of the battery, and connect the resistor and LED between the battery and the positive (+) lead of the capacitor as shown in **Figure 8A**. It should glow for 2 or 3 seconds, then gradually fade out. The capacitor is now charged — it won't accept any more electrons. Now, disconnect the battery, reverse the LED wires, and connect the resistor/LED combination across the capacitor while watching closely (**Figure 8B**). The LED will glow briefly as the energy stored in the capacitor flows through the LED and resistor until the charge is dissipated.

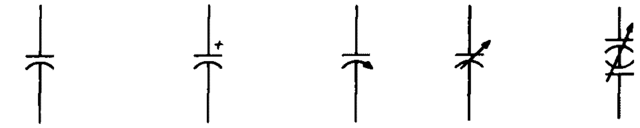
Do the same experiment with smaller values. Try a 5 or 10- μF value and note the difference in time. Connect several large ones in parallel and see how long they take to charge and discharge. To test the quality of the dielectric and insulation, charge a capacitor up and then wait several minutes before connecting the LED. If the capacitor makes the diode blink after 5 minutes, see if it still works after 10 or more minutes. Some capacitors have a very high quality dielectric and will store a charge for an hour or longer; others last only a couple of minutes.

What does the voltage rating mean?

You'll note that electrolytic capacitors have a voltage listed along with the capacitance. Unless otherwise stated, this is the maximum working DC voltage for which that type of capacitor is designed. You'll see many electrolytic capacitors for transistorized circuits rated for 6, 12, 15, 20, or 35 volts. You can always use less than the rated voltage without a problem. It's bad practice to use a capacitor right at its rated voltage because a voltage surge will short the unit sooner or later. It's a definite "no no" to use a voltage higher than the rating.

Nonelectrolytic capacitors, like disk ceramic, dipped mica, or molded

FIGURE 9



Schematic symbols for various types of capacitors.

paper have voltage ratings too. These units differ from electrolytic capacitors in that they can be reversed in the circuit; that is, they are not polarized.

Many nonpolarized capacitors have a black band around one end of the body, or have one end that is entirely darkened or colored to differentiate it from the other. This was called the "outside foil" in older molded paper capacitors. These capacitors were made by rolling up two strips of aluminum foil with a thin sheet of paper (often waxed paper) between. One foil was connected to one lead; the other was connected to the opposite lead. The layer of foil that ended up on the outside of the roll was designated as the layer to tie to ground because it helped shield the circuit from unwanted signal or noise pickup. If the circuit didn't go to ground, the outside foil was connected to the lower impedance or lower voltage part of the circuit. In modern circuits, even though most capacitors aren't made of rolled foil, there are still reasons to follow convention and designate one side as the "ground" or low impedance connection. In schematic diagrams, you'll notice that a capacitor symbol has one straight bar and one curved bar (see **Figure 9**). The curved bar is the outside foil side in nonelectrolytic capacitors and the negative side in electrolytic ones. Electrolytic units are further identified by a plus (+) sign near the straight bar in a diagram.

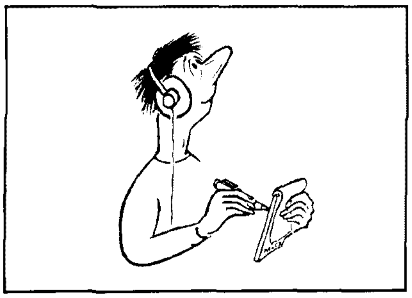
By Garth Stonehocker, KØRYW

RECEIVING DX BEACONS

Last month I presented beacon fundamentals, including beacon criteria and the frequency ranges of established beacons. As I mentioned, there are other radio stations that meet enough of the criteria to be useful as beacons. Among the stations that qualify are those owned and/or operated by Amateurs, broadcasters, researchers, and governments. Two examples of such stations are the Amateur network, PROPNET, on 14.1 MHz from worldwide locations and the standard frequency and time radio stations (also worldwide), mainly on 5, 10, and 15 MHz. There are also several useful stations on other frequencies. The best beacons would be from continuously on-the-air radio stations, identifiable not only by frequency but by voice or CW modulation, transmitting on several frequencies spaced across the HF frequency range (perhaps into the MF and low VHF ranges) and, of course, at locations known worldwide. Other information about the station like power and antenna (type and beam direction) would also be a plus.

Here's a method that will let you receive and use the ultimate beacon station, or those of lesser attributes. Radio Shack carries a crystal-controlled receiver for 5, 10, and 15 MHz plus VHF for NOAA weather; the price is $39.95. It doesn't have an RF input stage, so it's limited in sensitivity. Marrying it with an active antenna fixes the sensitivity problem and provides an omnidirectional indoor antenna to boot, or a connection for an external antenna if necessary. One such active antenna, the MFJ-1020A, is available from MFJ Enterprises, Box 494, Mississippi State, Mississippi 39762 for $79.95. Similar active antenna units have been featured in Amateur Radio magazines as homebrew weekend projects.[1,2] Because no metering is involved, this unit is only used for listening.

A high-tech beacon receiving setup can use a programmable receiver like the Kenwood R-5000, then do an analog to digital (A/D) conversion of its auto-



matic gain control voltage with an A/D board in the controlling microcomputer/PC. Software in the PC can execute control of the R-5000 at selected frequencies and times, control interrupts for the A/D functions, and manage the digital A/D data in storage and graphs. The graph would show the stored data and the latest values for a running plot of signal strengths at the monitored frequencies versus time. An automated system like this leaves you with little to do but view the graph for the beacon information you need.

There are, of course, all sorts of intermediate receiving setups for single or multiple frequencies. A hands-on application of a general coverage receiver (digital frequency dial and good image rejection would be advantages) using its S-meter for signal strength data values is one compromise. Another way to create a single frequency dedicated beacon receiver is by homebrewing. You can make an HF receiver/converter using a $3 pc board without the loopstick from a transistor AM radio with a three-transistor RF-mixer crystal oscillator. The converter circuit has been publised in Amateur magazines and the *ARRL Handbook*. Receiving beacon signals for propagation information can be an interesting part of DXing. Why don't you give it a try?

Last-minute forecast

The first two weeks of February are expected to have lower solar flux levels so the ionospheric MUF may be down, favoring the lower frequency bands. Signal strengths should be high and distortion from multipath lower. Thunderstorm noise is almost nonexistent and geomagnetic disturbances should be limited to the third and fourth weeks, possibly around the 16th and 23rd. These weeks are expected to exhibit higher solar flux levels, causing absorption and MUFs favoring the higher frequency bands with longer openings. No significant meteor showers are scheduled to appear in February. A full moon will occur on the 10th; perigee will be on the 7th.

Band-by-band summary

Ten and 12 meters, the highest day-only DX bands, are nearest (although somewhat below) the MUF on Southern Hemisphere paths. They'll be open most days when the solar flux is above 200 during the 10 to 14-hour period centered on local noon. These bands open on paths toward the east and close toward the west. The paths are up to 4000 km (2400 miles) in single hop length and, on occasion, double that during evening transequatorial openings.

Fifteen meters, a day-only DX band open most of each day, has lower signal strengths and greater multipath variability than 10 and 12 meters. It will be best when the MUF is just resting above this band and will remain so until it drops below the band — a transition period that occurs right after sunrise and just before sunset. Transequatorial openings will occur, with distances similar to 10 and 12 meters.

Seventeen, 20, and 30 meters are both day and nighttime DX bands. Seventeen is the maximum usable band for DX in the northern directions during daytime. In combination with 20/30 meters, it provides nighttime paths for the day-only bands. Thirty meters becomes the main over-the-pole daytime band, with some hours covered by 17 and 15 meters. This path may be affected by anomalous absorption on a few days of the month.

Forty, 80, and 160 meters, the night-only DX bands, exhibit short skip during daylight hours during the lowest solar flux, then lengthen at dusk. They are always far below the MUFs, except during disturbances on northern and east-west paths. These bands follow the darkness path, opening to the east just before local sunset, swinging more to the north-south near midnight, and ending up in the Pacific areas for a few hours before dawn. hr

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FEBRUARY

WESTERN USA

GMT	PST	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	4:00	17	30	12	10	10	10	10	12
0100	5:00	15	30	15	10	10	10	10	12
0200	6:00	15	30	15	10	10	10	10	12
0300	7:00	17	30	15	12*	10	10	10	15
0400	8:00	20	30	17	12	10	10	10	17
0500	9:00	20	30	17	15	10	10	12	20
0600	10:00	30	30	17	15	12	12	12	20
0700	11:00	30	30	17	15	12	12	15	20
0800	12:00	30	30	17	17	15	15	15	30
0900	1:00	30	30	17	17	15	15	15	30
1000	2:00	30	30	17	17	17	15	17	30
1100	3:00	30	30	20	17	17	17	17	30
1200	4:00	30	30	20	17	17	17	17	30
1300	5:00	30	20	10	12	20	17	17	30
1400	6:00	30	15	10	10	20	17	17	30
1500	7:00	30	15	10	10	10	17	15	30
1600	8:00	30	15	10	10	10	17	15	30
1700	9:00	30	15	10	10	10	15	15	30
1800	10:00	30	15	10	10	10	12	17	30
1900	11:00	30	17	10	10	10	10	15	30
2000	12:00	30	20	10	10	10	10	12	15
2100	1:00	30	20	10	10	10	10	10	12
2200	2:00	30	20	10	10	10	10	10	12
2300	3:00	30	30	12	10	10	10	10	12

MID USA

GMT	MST	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN	CST
0000	5:00	30	30	15	10	10	10	10	12	6:00
0100	6:00	30	30	15	10	10	10	10	15	7:00
0200	7:00	30	30	15	10	10	10	10	17	8:00
0300	8:00	30	30	15	12	10	10	15	20	9:00
0400	9:00	30	30	17	12	12	12	15	20	10:00
0500	10:00	30	30	17	15	12	12	15	20	11:00
0600	11:00	30	30	17	15	15	15	15	30	12:00
0700	12:00	17	30	17	17	15	15	17	30	1:00
0800	1:00	15	30	17	17	15	15	17	30	2:00
0900	2:00	15	30	17	17	17	17	17	30	3:00
1000	3:00	17	30	17	17	17	17	17	30	4:00
1100	4:00	20	30	20	17	17	17	17	30	5:00
1200	5:00	20	20	10	12	17	17	17	30	6:00
1300	6:00	20	15	10	10	20	17	17	30	7:00
1400	7:00	30	15	10	10	15	17	15	30	8:00
1500	8:00	30	15	10	10	10	15	15	30	9:00
1600	9:00	30	12	10	10	10	15	15	30	10:00
1700	10:00	30	12	10	10	10	12	17	30	11:00
1800	11:00	30	15	10	10	10	12	15	30	12:00
1900	12:00	30	15	10	10	10	10	12	30	1:00
2000	1:00	30	17	10	10	10	10	10	17	2:00
2100	2:00	30	20	10	10	10	10	10	15	3:00
2200	3:00	30	30	10	10	10	10	10	12	4:00
2300	4:00	30	30	12	10	10	10	10	15	5:00

EASTERN USA

GMT	EST	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE CARIBBEAN S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	7:00	30	30	15	10	10	10	10	17
0100	8:00	20	30	15	10	10	10	10	20
0200	9:00	30	30	15	12	10	15	15	20
0300	10:00	30	30	17	12	12	15	15	20
0400	11:00	30	30	17	12	12	15	15	30
0500	12:00	30	30	17	15	15	15	17	30
0600	1:00	30	30	17	15	15	17	17	30
0700	2:00	30	30	17	17	15	17	17	30
0800	3:00	30	30	17	17	15	17	17	30
0900	4:00	30	30	17	17	17	17	17	30
1000	5:00	30	17	10	17	20	17	17	30
1100	6:00	17	15	10	12	17	17	17	30
1200	7:00	15	15	10	10	12	17	17	30
1300	8:00	17	12	10	10	12	15	15	30
1400	9:00	20	12	10	10	10	15	15	30
1500	10:00	20	12	10	10	10	15	15	30
1600	11:00	20	12	10	10	10	17	15	30
1700	12:00	30	15	10	10	10	15	15	30
1800	1:00	30	15	10	10	10	12	15	30
1900	2:00	30	15	10	10	10	10	12	30
2000	3:00	30	17	10	10	10	10	10	17
2100	4:00	30	20	10	10	10	10	10	15
2200	5:00	30	30	10	10	10	10	10	12
2300	6:00	30	30	12	10	10	10	10	15

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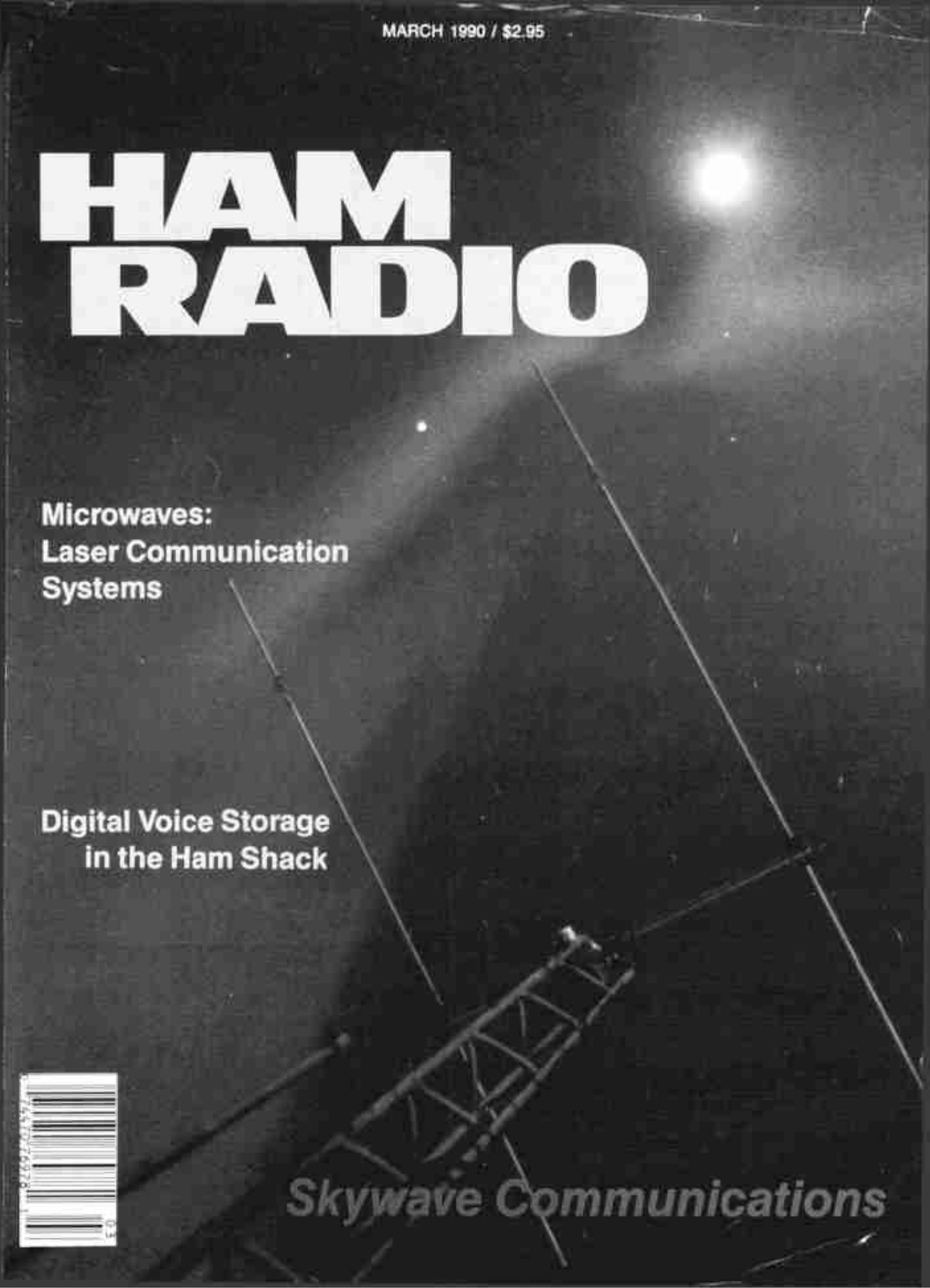
HAM RADIO

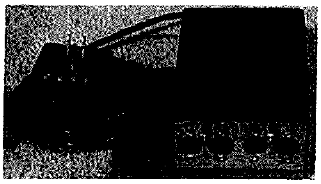


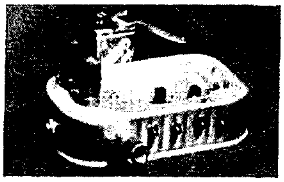
Microwaves: Laser Communication Systems

Digital Voice Storage in the Ham Shack

Skywave Communications

HAM RADIO


K3OF, page 22


WØJF, page 82

MARCH 1990
Volume 23, Number 3

Cover photo: Cornell Drentea, WB3JZO. Shot is of an Aurora Borealis over WB3JZO's QTH.

FEATURES

DEPARTMENTS

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Backscatter

Amateur Radio growth?

I understand that the running number of radio Amateurs remains nearly constant. That is to say, the population is not increasing. To combat this, a no-code VHF license has been proposed that presumably will remove a large stumbling block for the would-be ham. The result will be more radio Amateurs in the long run, making the Amateur Radio service more viable and healthy. Right?

Perhaps. However, we may be barking up the wrong tree. I suggest you take an hour or so and listen to the spectrum between 26 and 29 MHz. CB radio, as it was known, has disappeared and a new form of "hobby" radio has grown up in its place. The full 3-MHz range is chock full of SSB stations going about their business in a "ham-like" attitude. *Seemingly* absent are the loud-mouthed ignoramuses who made CB radio a shambles a few years ago. In their place are thousands of operators, behaving themselves, and having fun! There's plenty of DX in this range, too. I counted 14 countries in about 30 minutes listening time. Not much talk about equipment, but a lot of chatter about friendships and local color. It sounded very interesting. Too bad these thousands of operators are not hams!

But why should they be hams? What's the advantage? They can converse and enjoy themselves with no danger from the FCC. They exchange QSL cards and other pleasantries — and they have 3 MHz of space to do it! More frequencies than any HF ham band. The future radio hams are already on the air, and I don't see any chance of them becoming licensed Amateurs because there's little in it for them! They don't need Amateur Radio as we know it. A modified ham transceiver and a store-bought beam puts the operator on the air, ready to work DX and make new friends. To add insult to injury, I recently received a QSL from an English Amateur. On the card were his call letters and also his "identifier" for CB radio. He had the best of all possible worlds.

As long as the FCC and other licensing bodies allow unlicensed communications to take place in the 27-MHz region, we can't expect Amateur Radio to have much growth. The competition is too strong. And Amateur Radio shouldn't be blamed for either this problem or the lack of growth of the Amateur population! I'm not sure what the solution is, but I do know that a lot of happy people are enjoying the fruits of Amateur Radio the easy way. Can you blame them?

I propose that the ARRL initiate special broadcasts from W1AW in the 27-MHz range. This will, however, require special authorization from the FCC. These daily voice transmissions would include information about Amateur Radio, giving an address to send for more information on how to get a ham license. As time goes on, lessons in Morse code, rules and regulations of the Amateur service, and help with the Amateur exam could be given. A regular on-the-air course in Amateur Radio should be initiated and it could be broadcast right where it's *most* needed — in the middle of the "hobby" frequency range. I think that 1.5-kW PEP into a multi-element Yagi aimed at the heart of the country would gain a lot of attention among those who are potential radio Amateurs. Just as the Voice of America aims its broadcasts to selected areas of the world, the ARRL could aim its special 27-MHz transmissions at would-be Amateurs. Let's convince them that there is value in getting an Amateur Radio license!

Bill Orr, W6SAI

Comments

Creative recruitment; Productive leisure time

Dear HR

It is basic construction articles like W3RMD's ("The Five-Band Junkbox Transmitter," *Ham Radio*, December 1989 and "The 80/40-meter Junkbox Rig Revisiting, *Ham Radio*, January 1990) which, in my view, will go far to excite the imaginations of prospective hams and bring them into the fold.

Just as many of us got our feet wet with crystal control and "graduated" to additional frequencies and modes via VFOs and upgrading, so...might beginning with discrete components be the route for today's beginner to license upgrading and state-of-the-art technical skill and knowledge.

We Elmers with well-stocked junkboxes would do well to share our goodies with beginners, young and old, 807s anyone? How about an ARC-5?

Can we target as propective hams not only those with demonstrated technical skills and interests but also youth and adults who seek a leisure time activity that is not necessarily related to their intended or actual vocations or professions? I think such a strategy would be particularly relevant to older persons, whose numbers in our society are increasing dramatically and for whom Amateur Radio is an opportunity for continued active involvement in life and living that is service oriented. Three eminent hams come to mind in this regard — Senator Barry Goldwater, the late General Curtis LeMay, and King Hussein of Jordan.

Finally, can't we market CW as a second language to prospective hams rather than endure the frustration of defending/rejecting an arguably outmoded if not obsolete communications mode. Those of us who have mastered CW are, after all, bilingual.

Carlton D. Trotman, W3BRX, York Pennysivania

Relaying In Our Roots

Dear HR

What's all the "Bruhaha" about in regard to third-party traffic on the ham bands? As long as it doesn't violate any rule or regulation, national or international, what's the problem?

Looking back at the early beginnings of Amateur Radio, wasn't the "relaying of messages" *one* of the very many public services that the Amateur Radio operator helped develop, and very efficiently at that? This activity was instrumental in promoting the founding of the ARRL (American *Radio Relay* League).

As for self-policing, no group or society can achieve successful results with that method alone.

We still need a *strong* FCC which needs all the help we can offer; i.e., official observers.

Alex Hellman, W2OEQ, Woodhaven, New York

AM alive and well!

Dear HR

This is in reference to "Ham Radio Techniques," by Bill Orr, W6SAI, appearing in the December 1989 issue of *Ham Radio*. We take issue with the statement: "Too bad the days of amplitude modulation are past..." It's a matter of fact that 29.0 to 29.1 MHz is populated with those using AM. Listen to 14.286 MHz just about anytime and you will find it quite busy. Good 'Ole 160 is still the local favorite for the lower powered AM signal; we use 1890 here in Hawaii. Also, KH6CC and KH6B can be heard often on 7290 kHz using AM. We call your attention to the publication, *The AM Press/Exchange*, published by K4KYV. Also, *Electric Radio*, put out by N6CSW/Ø. No, AM is far from dead! We have heard that it's the fastest growing mode. Many modern solid-state rigs have an AM mode position (wonder why?). Those who, for the first time, heard Amateur AM signals have switched their rigs over to AM and have noticed the "quality mode." Bill Orr, W6SAI, has had kind words towards AM in the past. See "Ham Radio Techniques" in the February 1984 issue of *Ham Radio*.

To conclude, we would like to state that many of us started in Amateur Radio with low powered rigs on 160 meters. Among them are KH6CC, KH6B, and we've heard that W6SAI (under his first call sign) spent a lot of his teenage years on 160 AM!

Jack Wheeler, KH6CC and Dean Manley, KH6B, Hilo, Hawaii

Practical and thorough...

Dear HR

Please let me report that John Piunichny's article on the dual eccentric capacitor drive ("Near Linear Tuning with Dual Eccentric Pulleys," January 1990) is the best article I've seen in the ham radio journals for a good long time. Let me also suggest that articles of this practical value and thoroughness be printed as often as possible. I hope we see more contributions by Mr. Piunichny in the future.

J. A. Smith, Hudson, Wisconsin

SKYWAVE COMMUNICATIONS PART 1

A brief account of the propagation phenomena

By Cornell Drentea, WB3JZO

"To account for the transmission of waves through space containing no ordinary matter it seems necessary to assume the existence of a universal medium filling all space and even interpenetrating matter itself."
From *A Text Book of Physics*, by A. W. Duff, circa 1908.

The ether

If radio signals were transmitted in free space regardless of their frequency, they would appear to propagate along straight lines, only to be curved by time itself within the theory of quantum electrodynamics. Within this concept, waves from a transmitter's antenna (providing that a true omnidirectional pattern could be obtained) radiate an electromagnetic field in the entire space surrounding it. In this ideal case, a receiving antenna in its field would receive a certain amount of the transmitted energy, which would be inversely proportional to the distance from the transmitter. That is, power flux degrades inversely as the square of distance from the transmitter point. Propagation would then be defined as the transfer of energy without the transfer of matter. This is the modern theory of free space propagation.

However, this theory wasn't always known. Until recently, scientists believed that in order for electromagnetic waves to propagate, there had to be a medium (versus a vacuum) — just as air or some other medium is needed for mechanical sound propagation. For some time, it was believed that a certain exotic substance called "ether" filled the universe. The ether theory came about in 1865 when the British physicist James Clerk Maxwell described mathematically how wavelike disturbances in the combined electromagnetic field would travel at a fixed speed. The speed of light was first measured in 1676 by the Danish astronomer Ole Christensen Roemer. He observed differences in events happening with Jupiter's moons which led him to the conclusion that light travels at a finite speed. Imprecise measurements taken at the time indicated that this speed was 225,308 km per second (or about 140,000 miles per second) rather then the modern value of 300,000 km per second (or about 186,000 miles per second).

Maxwell showed that if the distance between the wave peaks (the wavelength) is a meter or more in length, you have radio waves. Consequently, shorter wavelengths would be known as microwaves, infrared, visible light, ultraviolet, x-rays, and gamma rays.

Maxwell's theory was very advanced and accounted for the present theory of relativity which says, among other things, that light travels at a finite speed. However, at the time, an older Newtonian theory existed which said that the speed of anything was to be measured relative to something fixed — just like the speed of sound in the air.

It was then suggested that electromagnetic energy was propagating through the mysterious substance, and that the speed of electromagnetic energy was measured against this fixed ether. As a result, different observers moving relative to the ether on two separate points on earth would perceive light coming toward them or moving away from them at different speeds, but the light speed relative to the ether always remained fixed. According to this theory, as the earth was turning around its axis and floating in ether, the apparent speed of light from a single source was higher or lower — depending upon the location of the observers. But experiments carried out in 1887 by Albert Michelson and Edward Morley at the Case School of Applied Science in Cleveland, Ohio indicated that the speed of light observed in the direction of earth's rotation was exactly the same as that at right angles to the earth's motion.

This defeated the mechanical wave propagation in the air model and, of course, the ether theory. It was proven soon after, in a famous paper written in 1905 by a clerk named Albert Einstein of the Swiss patent office, that the

whole theory of ether was unnecessary if one was willing to abandon the idea of absolute time. Thus, the theory of relativity was born. Einstein's theory said simply that the laws of physics should be the same for all free moving observers, no matter what their speed. Since then the term ether has been used only to recall our naive vision of the universe during those beginning radio days. (Our knowledge today may still be naive in view of what we will learn tomorrow.)

Electromagnetic wave propagation on earth

I have discussed how, in the ideal propagation model, a space-transmitted RF signal leaves the antenna and propagates in all directions without the help of ether. This energy would be received at a point located on an infinite number of imaginary spheres surrounding each other like the layers of an onion. There is a big departure from this simple concept when you look at what actually happens to wave propagation on earth. First, these ideal spheres are broken up by the earth's mass and magnetic field, so the ground may indeed act as part of the transmission circuit. This interaction usually results in losses which diminish the groundwave with distance. We know today that different ground conditions provide different degrees of attenuation. For example, over a good conductor like seawater, absorption tends to be minimal for very low frequency (VLF), low frequency (LF), and some medium frequency (MF) waves — a condition used to enhance maritime services ranging from 10 to 100 kHz.

In addition, depending on the frequency of an RF signal, complex and not entirely predictable interaction with the composition of the earth's atmosphere exists due to the sun's indirect meteorological impact. Things are further complicated by the interaction of the sun's radiation with the matter in the atmosphere. This is known as the ionospheric phenomenon and is further impacted by the earth's magnetic field. Propagation of high frequency waves on earth is therefore impacted, in addition to the free space loss discussed earlier, by the ionosphere absorption, dispersive loss, and by ionospheric focusing and defocusing.

The short path and the long path

The shortest possible communication route between any two points on earth follows a curved path around the earth between the two points. This is known as the short path. The line is part of an imaginary circle drawn around the earth's curvature, called the great circle line. This line touches the two points and is an extension of the short path. The longer route between the two points on the great circle line is known as the long path. Finding the great circle line of any two points on a globe is easily achieved by stretching a rubber band around the globe and touching the two points. With the earth's circumference of 38,624 km (24,000 miles), there will be only one instance in which the paths are equal; that is, when the two points are 19,312 km (12,000 miles) away from each other. In all other instances, there will always be a short and a long path. In general, the short path is used for communications, but not necessarily via direct waves. Long path communication occurs if low angle ionosphere reflections and refractions exist, and there is evidence that both paths can be engaged at the same time in high frequency (HF) communications. The propagation via the two paths and the ionosphere is defined as skywave communication.

The challenge of predicting propagation

In general, HF radio communication depends upon the ability of the ionosphere to reflect the transmitted radio energy back to earth over the short and long paths. Although this is a well-known concept, predicting the degree of ionization in the various regions of the ionosphere, anticipating the expected skywave signal at any point on earth, and comparing this information with the expected local radio noise environment at a particular time, have been the subjects of much work intended to prove the *reliability* of radio circuits. This work has also had an additional impact on the design and development of radio communications receivers.

The prediction of ionospheric performance has been based mostly on empirical work. The problem lies in properly defining ionospheric absorption equations and combining them with the theoretical ground loss, dispersive loss, free space loss, power focusing and defocusing, and antenna gain factors. This is further aggravated by the effects of solar activity, and the seasonal and diurnal variations.

Despite all these difficulties, computer models of parabolic distribution (the ionosphere looks like a concave mirror from earth) of electron density have been developed along with new concepts. In addition to the great circle line-single ray concept discussed earlier, these concepts concern such new things as complex azimuth diversity and coverage by stochastic (refers to independence of events of random variable nature) scattering phenomenon.

The art of HF propagation prediction has been furthered by field tests and backed by a worldwide network of vertical incidence ionosondes intended to measure the diverse parameters at heights of up to 300 km (186 miles). Additional work has been performed at Arecibo Observatory with a new technique known as incoherent scatter observation (observation of the scattering of the transmitted waves by the individual electrons in the ionosphere). This technique allows scientists to observe the ionospheric conditions to heights of about 1,995 km (1,240 miles). Heating and artificial excitation of the ionosphere have also been done through the observatory's ionospheric modification facility (nicknamed "the heater") near Arecibo. This facility has 58 acres of log-periodic antennas powered by an 800-kW transmitter which can concentrate energy in the atmosphere at wavelengths of 20 to 200 meters. The heated electrons and their interaction can then be studied by the 1,000-foot radio telescope. Worldwide radio noise measurement records have been compiled since 1963 in a famous report entitled "World Distribution and Characteristics of Atmospheric Radio Noise."[1]

Despite all these advances, the science of propagation prediction remains an inexact one. In part 2, I'll look at what is known about the ionosphere.

REFERENCE

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The Night of the Aurora

It was a beautiful clear night in October 1989. The Indian summer was still with us here in Lake Wobegon country, and a full moon was majestically displayed over the cold evening sky. The smell of leaves made me think of the Minnesota winter just around the corner. Suddenly, I felt isolated from the rest of the world and overcome with a strong desire to search the ether for other lonely beings. Tuning across the bands produced a couple of short contacts, but signals were faint and rubbery — like weak, fluttering voices at the end of a tunnel. I could still hear a faraway storm crushing its way through the south. And then...it all turned to silence There was nothing but a slowly growing noise which covered every band. My digital clock showed 3:17:30 a.m. when I decided to listen to the propagation bulletin just coming up on my WWV receiver. But I couldn't hear anything there either. I thought this must have been the way the bands were before the days ot radio, filled only with pulsating noise which took over everything. Through the window I could now see a growing purple light, which at times seemed brighter than the light coming from the moon. "It must be Aurora Borealis," I determined as I turned off all radios and rushed out into the yard. And there it was: majestic, right above my head with long purple drapes extending all the way to the ground; it seemed I could almost touch them with my bare hands. I had an overpowering electrical feeling which I have experienced only once before, while watching fireball lightning at sea during an electrical storm in the tropics. The light was waving slowly and changing colors. Then it stopped for a moment, only to start moving again.

I rushed in for my camera and tripod. (The result appears on the cover of this issue. Ed.) Like any picture, it can only express a fraction of the original feeling I experienced when I took it. I wrote this story in an effort to share the experience The work which accompanies it is intended to further your knowledge about the sun and ionosphere, and how they affect radio communications. Although much has been written on the subject of propagation, this three-part article will stick with the scientific base. I intend to clarify the concepts of HF communications rather than chronologically list data. Because the subject is so vast, I chose to emphasize certain technical areas that cannot usually be found in one single publication, and combine them with facts and experiences of my own. The work begins with the free space propagation concept, concentrates on complex ionospheric physics made simple, and concludes with a discussion on tools for predicting propagation. In presenting the propagation software at the end of part 3, I do *not* make any claim as to performance, nor am I an agent of any of the companies involved. Anyone interested in obtaining more information about this software should contact the producers directly, or obtain the information from advertisements.

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ANOTHER LOOK AT HF MOBILE ANTENNAS

By Jack Najork, W5FG, 723 Flamingo, Duncanville, Texas 75116

After reading NC0B's excellent article on HF mobile antennas in *Ham Radio*'s September 1989 issue,[1] I was inspired to look up some data I recorded many years ago on inductively loaded mobile antennas. I'd like to offer some additional thoughts on this subject.

Loading coils

It has long been stressed that HF antennas requiring inductive loading should use high Q, low loss coils. This concept is employed in the "bug catcher" type coil. The coil is perhaps 3 to 5 inches in diameter and space wound with no. 12 or no. 14 wire to give the optimum form factor for maximum Q. A length-to-diameter ratio of 0.4 to 0.5 results in the shortest length of wire and, therefore, the lowest RF resistance for a given inductance.

While this form of loading coil is generally an improvement over some commercial designs, it's not the optimum when used as part of the overall mobile radiation system.

Tests indicated that the Q factor of the coil is secondary to its *physical shape.* A long, narrow coil (still low loss) of lower Q consistently produced a stronger radiated signal than the short, fat, high Q coil when all other factors of the system were equal.

Shape versus Q

One theory for the superiority of the long coil was given by E.L. Gardiner, G6GR, in *Radio Communication.*[2] He pointed out that any radiated field in space must have both an electrostatic and an electromagnetic field which are correctly related. Neither field by itself will produce meaningful radiation.

In the typical loaded mobile whip, current in the lower section generates a magnetic field. This field won't be radiated unless an adequate electrostatic component is also present in the form of an RF potential difference between the ends of the conductor carrying the current; that is, the base and tip of the whip. These components will be in phase because the antenna is a resonant circuit. The major portion of the potential difference appears across the ends of the coil, as is normal in a parallel-tuned circuit. The electrostatic field strength setup is proportional to the distance between these two high potential points — namely, the length of the coil.

For example, 100 volts across 1 meter represents an electrostatic field of 100 volts per meter, while the same potential across 1 centimeter represents only 1 percent of this field. This leads to the conclusion that however strong the electromagnetic field component may be, it can be fully transformed into radiation instead of heat only if an adequate electrostatic field is present, and vice versa.

Practical considerations

From a practical standpoint, the optimum length for this type of loading coil is 14 to 20 inches for 160 meters and 10 to 14 inches for 75 meters, with corresponding diameters of 1 to 2 inches. The lower wind resistance of such a coil allows it to be located higher up on the vehicle. This places the radiated field away from the car body and reduces losses.

For maximum radiation and efficiency, the loading coil should be placed as high above the vehicle as possible. I have seen bug catcher type coils mounted on vans with a spacing of a foot or so from the metal body — a sure method of converting most of the RF to heat instead of radiation.

It isn't difficult to homebrew efficient mobile antennas. You can use any sturdy material for the bottom section. I use fiber glass wound with copper tape. I also use fiber glass for the top section. My loading coil forms are polyethelene bottles of the desired diameter. I remove the top and bottom of each bottle to produce a hollow form. I use foam tape on the fiber glass whip to build it up to a diameter which is a snug fit for the top and bottom of the form. For 75 meters, you may need two forms situated one above the other. After winding, cement the coil to the taped segments. Because the fiber glass whip passes through the coil, it receives very little physical stress. This method also eliminates the need for any large metal fittings near the field of the coil which would reduce Q. The only metal parts near

the coil are two 4-40 brass bolts and nuts used to anchor the ends of the winding.

Coils are space wound with no. 18 wire for 75 meters and no. 16, 14, and 12 for progressively higher frequency bands. If you use fiber glass for the complete top section, make the segment conductive by slipping on braid from coax cable. Because you want maximum capacity above the coil, cover the entire surface of this section with the braid. A single piece of wire (no. 14, for example) taped to the fiber glass will work but the capacity will be lower, requiring more inductance on the coil.

After you prune the coil to frequency, give it a layer of PVC tape for weatherproofing and to eliminate "bug catching" between turns. Drill several holes through the layers of tape at the bottom to let the inside of the coil breathe and to prevent moisture buildup.

Please note that this is a single band antenna. I change bands by changing top sections, each of which is optimum for one band. The inconvenience of changing top sections is offset by the "home station" reports I receive.

Tuning the antenna

Commercial mobile antennas are generally close to resonance and require only slight adjustment of the top section to bring them on frequency. Homebrew antennas, however, can initially be megahertz away; this can pose a tune-up problem.

The usual method of finding resonance involves coupling a grid dip oscillator (GDO) to a one-turn loop connected between the base of the antenna and car body (ground). You then find the frequency of the GDO on a calibrated receiver.

If you don't have a GDO, you can get the same results with a 79-cent Radio Shack buzzer and some flashlight cells. Couple a loop from the buzzer-battery connection to the loop at the base of the antenna. Temporarily shield the buzzer by sticking it in a cake tin or wrapping it in foil. With the buzzer fired up, the antenna will be shock excited and radiate a weak, raspy signal at its resonant frequency. You'll be able to pick up this signal on your receiver. If the band is busy, you may need to run a piece of coax from the receiver to the vicinity of the antenna to pick up antenna radiation only. This system works best on the lower frequency bands where the Q of the antenna is higher, with buzzer noise peaking over a 20 to 30-kHz segment of the band.

After you fracture your rib cage laughing at this scheme, I'll tell you that I filched the idea from the broadcast industry. In the days before fancy instruments, the same system was used to determine the resonant frequency of scaled broadcast towers. If your library is as old as mine, you'll find this technique in *The Radio Engineering Handbook* by Keith Henney.[3]

After you determine the resonant frequency (hopefully at a lower than desired frequency) remove turns from the loading coil, one at a time, until you observe resonance on the high frequency end of the band you want. Turns must be removed, not shorted; a shorted turn will lower the Q of the coil.

On 160, 75, and 40 meters, it's easiest to retune to lower frequencies with a remotely controlled roller coil in the car trunk, as described by K9MLD in *Ham Radio*, October 1988.[4] I used this system back in the fifties running 35 watts AM on 75 meters and can vouch for its convenience. The secret of efficient operation is to peak the antenna alone at the high frequency end of the band and then use just enough roller coil inductance to restore resonance on lower frequencies. Because the roller coil is bottled up in your trunk, any radiation it produces isn't going anywhere, so use as little inductance as possible. The best resonance indicator for this system is a field strength meter at the driver's position; this will quickly tell you when the system is peaked to maximum output.

Some homebrewers use a sliding section type whip above the coil to tune. Unfortunately, the conventional auto type collapsible whip will quickly develop intermittent contact at the sliding joints, resulting in noisy reception and erratic loading on transmit. Sliding joints are recommended only if you devise some way of fastening them securely after adjustment via a set screw to ensure low loss continuity. Even a few watts of RF power will affect joints in this manner — they just aren't designed to pass RF current.

In lieu of a roller coil, you can make limited excursions lower in frequency using an alligator clip and a short (2 to 3 inch) piece of wire clipped above the loading coil to increase capacity. After a few trials, you can readily determine where to locate the clip in order to hit the desired band segment.

SWR

No story on antennas is complete without a discussion of SWR. I may shock the majority of you by saying that the usual obsession with a low SWR doesn't really apply to mobile HF antennas. Other than to satisfy a fussy solid-state rig, a low SWR isn't essential. In some cases it's actually detrimental.

Your objective is to obtain maximum radiation from an antenna system which is, by its physical properties, relatively inefficient. The fallacy of striving for a low SWR lies in the fact that you can doctor the antenna to produce a low SWR and, in so doing, actually reduce the effective radiation of the system. The best method of tuning a mobile antenna is with a field strength meter, adjusting for maximum radiation. Once you've done this, you can check the SWR. On the lower frequency bands it can be 2:1, or even 3:1. Because most transmission line runs on mobiles are short (less than 20 feet), the losses incurred from such ratios are negligible. If your sold-state rig doesn't like this condition, a simple "L" network at the rig will make things right.

It's entirely feasible to use inductive or capacitive matching devices at the feedpoint (base) of the antenna to improve the SWR. However, these are generally one-band devices requiring readjustment for each band. So, if you're an SWR fanatic, you can use base matching or an "L" network. But remember that each requires attention for a band change.

The interstate bonus

For the crowded 20-meter band, I built two top sections. For in-town use I have a shorter section with a total height of 9 feet. On the interstate, where most overpasses are at least 14 feet, I go to a longer top section with the loading coil 10 feet above ground and a total height of 13 feet. This begins to look like a full quarter wave, and with my 100-watt homebrew mobile, performance begins to approach home station efficiency. The real interstate bonus comes

courtesy of the highway's construction. With their large masses of steel reinforcing bars under the concrete, these highways constitute an excelleht ground system which improves mobile operation significantly. It's the next best thing to driving over salt water (using a bridge, of course), so don't be surprised to find that signals drop suddenly when you leave the interstate for a country road!

You can always improve your mobile antenna system by becoming a fixed mobile. When parked at an off-the-road site, clip a quarter-wave length of wire to the *bottom* of the loading coil and throw the other end up into a nearby tree. This is most effective on 75 and 40 meters where antenna efficiency is lowest.

One ham I know kept an important 40-meter sked with this system. Unfortunately, he was out in west Texas and there were no trees around, so he had his XYL hold up the far end of the wire with his fishing pole. Needless to say, the QSO was short. hr

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2. E.L. Gardiner, G6GR, "The Practical Design of Mobile Aerials, *Radio Communication*, RSGB, July 1971, page 450
3. Keith Henney, *The Radio Engineering Handbook*, 4th Edition, Chapter 14, page 613
4. Joel Eschmann, K9MLD, "The Weekender: Remote Tuner for 75-Meter Mobiles," *Ham Radio*, October 1988, page 36

Microwaves

Bob Atkins, KA1GT

LASER COMMUNICATION SYSTEMS

Over the last year or two it seems that a growing number of Amateurs have been building laser communication systems. This is probably the result of two factors. First, the availability of surplus lasers is increasing while the cost is decreasing. Second, many of the VHF/UHF/Microwave contests now award extra points for laser contacts. Last fall at the Mid-Atlantic States VHF Conference I gave a talk on laser communications which seemed to generate a lot of interest, so I thought I'd cover laser communication systems in my first few columns. Understanding how to build and operate such systems requires knowledge of three factors: laser transmitters, laser receivers, and atmospheric effects on laser propagation. While a number of complex heterodyne laser communication techniques are possible, they are out of the realm of Amateur operation, so I'll deal only with simple direct detection systems here. This month I'll discuss the laser transmitter end of the link.

I think a little historical background would be useful. In 1917 Einstein postulated — as part of his theory of blackbody radiation — that when an atom in an excited state was hit by a quantum of radiation, it could be induced to emit radiation with the same frequency, phase, and direction as the incident quantum. The original incoming quantum of radiation isn't absorbed and amplification has been achieved because one radiation quantum has now become two. This process is known as the stimulated emission of radiation. The first practical use of this process came in the invention of a microwave amplifier in the early 1950s. This device was the MASER, which stands for Microwave Amplification by Stimulated Emission of Radiation. Later, the same principle was applied to optical frequencies and the optical maser was developed. This is now called the LASER, and stands for Light Amplification by Stimulated Emission of Radiation. Most lasers are used as light sources rather than amplifiers by using positive feedback (more on this later).

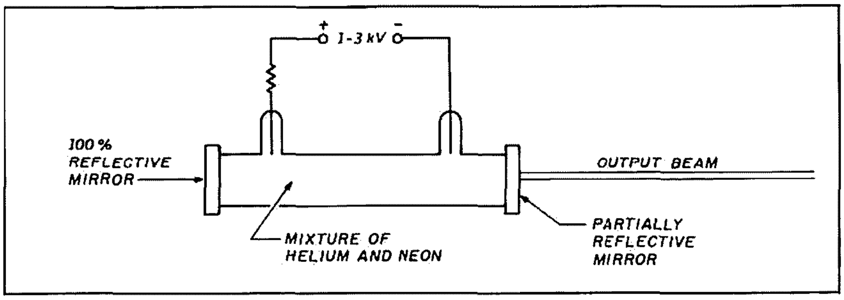
The many different types of laser are usually classified by the nature of the lasing medium. This medium can be a gas, liquid, or solid. The lasers which show up on the surplus market are almost always gas lasers. A mixture of helium and neon make up the lasing material. They are known as helium-neon or He-Ne lasers. Semiconductor lasers also show up from time to time, but they are generally less useful for DX communications purposes, so I won't deal with them here. The basic construction of a typical He-Ne laser is shown in **Figure 1**. A hollow glass or ceramic tube is closed off at each end by mirrors and filled with a mixture of helium and neon. One of the mirrors is 100 percent reflective; the other is only partly reflective, allowing some light to pass through. An electrical discharge is then set up in the tube, exciting some of the neon atoms. A few of these atoms emit light quanta through a process known as spontaneous emission. These light quanta can then collide with other excited atoms, causing stimulated emission (the amplification process) in the same direction as the incident quantum. Because there are mirrors at the ends of the laser tube, light quanta are reflected back and forth along the tube making many collisions with excited atoms in the process. This further amplifies the light and sustains the emission process through positive feedback. A small amount of light leaks out through the partially reflective mirror; this is the output laser beam. The wavelength of the output light is determined by the composition of the gas in a laser and the design of the tube and end mirrors. He-Ne lasers are normally designed to emit red light at a wavelength of 632.8 nm, but they can be designed to emit green light or even infrared radiation at lower efficiency. Other gas lasers, like the helium-cadmium (He-Cd) and argon (Ar), emit light mainly in the blue and blue-green regions of the spectrum. Note that the nature of the light output by a laser is usually characterized by its wavelength. A number of units are commonly used. They are:

Nanometers = 10^{-9} meters
Angstroms = 10^{-10} meters
Microns = 10^{-6} meters

The frequency of the output beam is rarely, if ever, used. Thus the red beam from a He-Ne laser can be characterized as one of the following, all of which are equivalent:

632.8 nm (nanometers)
6328 Å (angstroms)
0.6328 μ (microns)

FIGURE 1



Simple He-Ne Laser.

Expressed as a frequency these become:

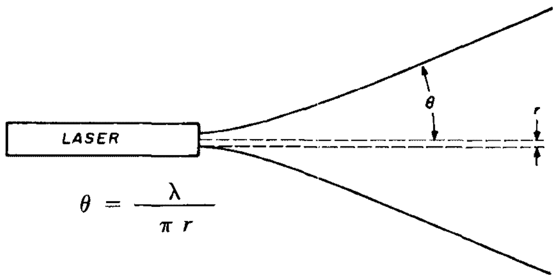
474.35 THz
474,350 GHz
474,350,000 MHz

Most of the He-Ne lasers on the surplus market have a power output of 1 to 5 mW, which is quite adequate for even long range DX contacts. Prices range from $40 to $200 depending on power output and condition.

A laser beam has several properties which distinquish it from other light sources. It consists of light of a single wavelength (or a very narrow wavelength range), in contrast to light from a flashlight which emits light over a very broad wavelength range. For an RF analogy, you might equate the laser with a single frequency crystal-controlled carrier, while the flashlight would be analogous to the output from a noise diode or even a spark transmitter! A second important characteristic of the output beam from a typical gas laser is its very small divergence (beam spreading). Even at a distance of about a mile, the beam from a small He-Ne laser spreads only to a diameter of 5 feet. A third unique feature of laser light is that it is coherent; that is, every light quantum, or photon is emitted in phase. This is very important in certain laser applications, like holography, but isn't a requirement for efficient DX communication.

From the standpoint of DX communication, perhaps the most important feature of the laser is the low beam divergence. The laws of physics indicate that all beams diverge, no matter how perfectly collimated (parallel) they are to start with, as a result of diffraction. The degree of divergence is directly related to the beam diameter — the larger the beam, the lower the divergence. A radio analogy can be found in parabolic antennas. A very large antenna produces a very wide initial beam with a very small beamwidth (divergence), whereas a small antenna produces a small initial beam with a large beamwidth (divergence). You can also look at this as a consequence of diffraction. The same equations govern both the spreading of a laser beam and the beamwidth of a parabolic dish. The geometry of diffraction spreading is shown in **Figure 2**.

FIGURE 2



$$\theta = \frac{\lambda}{\pi r}$$

where θ is the beam divergence in radians. (1 radian ≈ 57.3°)
λ is the wavelength
r is the beam radius
(λ and r are measured in the same units.)
For a He-Ne laser with a 1-mm diameter beam.

$$\theta = \frac{6328 + 10^{-7}}{(\pi) + (0.5)} = 4.03 + 10^{-4} \text{ Radians}$$

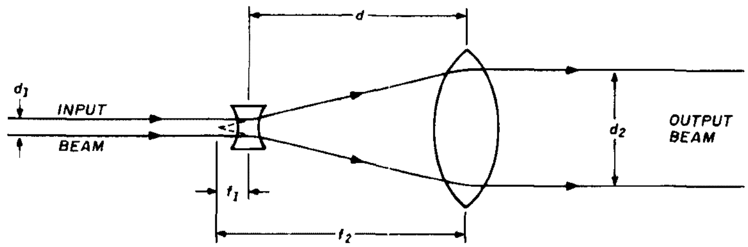
Total beam divergence = 0.806 mR or 0.046°

Diffraction limited beam spreading. Both θ and r are measured at the point at which the beam has an intensity of

$$\frac{1}{e}\left(\frac{1}{2.7}\right)$$

the intensity in the beam center (assuming a Gausian beam). θ then corresponds to the 4.3-dB beamwidth of the beam.

FIGURE 3



f1 = (negative) focal length of diverging lens (input)
f2 = (positive) focal length of converging lens (output)
Beam expansion = f1/f2
Lens separation = d = f2 − f1
With an input lens of focal length (−) 1 cm
and an output lens of focal length 4 cm:
The beam will expand by 4×.
The lens separation will be 3 cm.

Laser beam expander.

Increasing the beam diameter of a laser is analogous to using a higher gain antenna. Optically this is accomplished by means of a beam expander, as shown in Figure 3. This is akin to a Gallilean telescope used in reverse; that is, the beam enters through the eyepiece and exits through the objective. While such a decrease in beam divergence may be desirable for DX communication, it is not without significant problems. Most small He-Ne lasers have an intrinsic beam divergence of about 1 milliradian (1/20 degree). This means that the laser must be pointed at a distant receiving station with an accuracy of better than 1/20 degree. If the beam is expanded five times, the beam divergence drops by a factor of 5, and the required pointing accuracy becomes 1/100 degree. Obtaining such a pointing accuracy isn't an easy task. It requires a very solid mounting system and a capability for very fine positional adjustment, not only in azimuth but also in elevation. You can achieve this by using a system like the one shown in Figure 4. You can also try adding a sighting telescope as an alignment aid.

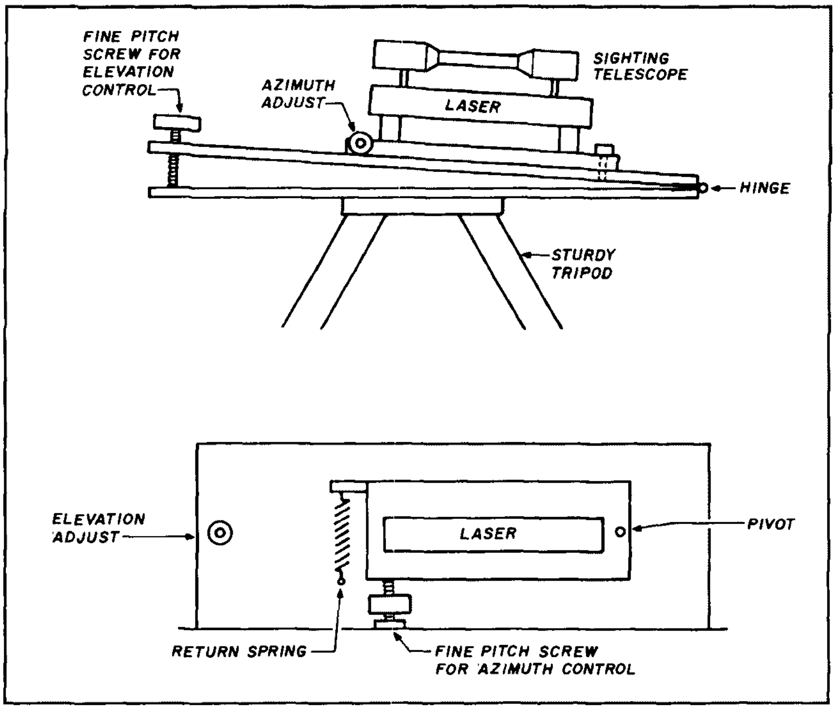
In order to transmit information via a laser beam, you must achieve some form of beam modulation. Mechanically interrupting the beam is the simplest, cheapest, and most efficient modulation scheme. At reasonable CW speeds you can do this using a solenoid operated shutter. This would correspond to A1A emission using the WARC '79 scheme. Alternatively, you can modulate the beam at an audio frequency by passing the beam through a rapidly rotating wheel with slots cut in it (try the blades of a fan). This modulated beam can be keyed on and off mechanically. This is modulated CW (MCW), or A2A modulation, under the WARC '79 designation. The advantage of using MCW is that it lets you use a simpler receiver system which I'll describe later.

If you want analog (voice) or high speed digital modulation, there are a couple of ways you can accomplish it. The preferred method is to use an acousto-optic modulator. This device is made up of a special type of crystal which is acoustically modulated at a very high frequency (several MHz) by the application of an RF field. The process is similar to the piezo-electric effect exhibited by quartz crystals. This acoustic modulation sets up standing pressure and density fluctuation in the crystal which can diffract a laser beam passing through it. As a result, a single input beam is diffracted into two (or more) output beams. The relative intensity of the power in the beams is a function of the RF modulating power. When the RF power is modulated, the output beams are amplitude modulated. While this method is capable of efficient and rapid analog or digital modulation (with bandwidths in excess of 1 MHz), its disadvantages are high cost (new acousto-optic modulators are $500++) and the complexity of the drive electronics. Surplus equipment (from laser printers or FAX machines) containing acousto-optic modulators is sometimes available on the surplus market for $50 to $100.

A second method of amplitude modulation involves modulating the high voltage supply to the laser. This isn't very efficient because the maximum amplitude variation in the output laser beam is about 15 percent (typically it's much lower), but it can be done inexpensively. The reason for the low modulation amplitude lies in the discharge process in the laser tube. It operates somewhat like a common neon bulb. A certain voltage is required to "strike" the discharge, but if the voltage is too high, the resultant high current will destroy the tube. Because the tube voltage must be held within quite tight limits, the resultant power output doesn't change greatly. The voltage modulation can be achieved by connecting one side of a well-insulated transformer in the lead carrying the high voltage to the laser tube. The other side of the transformer is then connected to an audio modulation source. You may need to experiment to find the optimum conditions for maximum laser modulation amplitude; some lasers may be more amenable to this type of modulation than others. Typical modulation levels will be on the order of a few percent.

There are numerous other modulation methods. These include: reflecting the laser beam from a small mirror attached to a loudspeaker, using transmission type liquid crystal displays (LCDs) as a shutter, and various kinds of magneto-optic and electro-optic devices (Kerr cells and Pockels cells). Though all of these methods can be made to work, they will generally be less convenient and less efficient than

FIGURE 4



Azimuth-elevation mount for laser. Note: The greater the pivot-azimuth screw and hinge-elevation screw distances, and the finer the screw pitches, the better the fine adjustment capability. Never point laser at any person, animal, or vehicle.

beam interruption or the use of an acousto-optic modulator.

The small, 1 to 5 mW, He-Ne lasers found on the surplus market require power supplies which produce 1 to 2 kV at a current of a few milliamps. Because the current drain is so low, you can build a fairly simple power supply using diode and capacitor voltage multipliers. For portable use, try using a transistorized DC inverter to generate an AC output of a few hundred volts which can then be multiplied up to the required voltage. WA6EJO described such a power supply in the December 1986 issue of *Ham Radio*. Very compact encapsulated power supplies are commercially available. A power supply for a 2-mW laser which runs off 12 volts DC would be about 3 × 0.75 × 1 inches and would draw about 1 A. You can find commercial supplies running off 117 volts AC for around $35; 12 volts DC supplies are more expensive at around $90 and up.

While most of my comments have dealt with He-Ne lasers, other types of laser are also suitable for communications use. In fact, KY7B and WA7LYI used He-CD (blue) lasers to set a DX record of 95 miles in August 1988. I've concentrated on He-Ne lasers because they are by far the most common and least expensive lasers found on the surplus market. It's also true that red light will suffer less attenuation than blue light in its passage through the atmosphere. (I'll discuss this further in a future column.)

The final, and perhaps most important, point to be made about lasers is that they can be dangerous. Laser power supplies produce high voltages. Even though only low currents are involved, such supplies can be lethal. The output beam, though low in total power, has a very high power density. When focused through the lens of the eye onto the retina it can easily destroy tissue, causing partial or total blindness. The degree of danger is a function of the laser power. Lasers of less than 1 mW are generally safe, and even direct exposure to the beam isn't likely to be harmful. At power levels of 1 to 5 mW, the light is so bright that there's a natural tendency to blink and turn away from it, thus avoiding damage to the eye. At power levels of 10 mW and above, the blink/aversion response is often not fast enough to prevent injury. Remember that these are average effects in normal adults. Injury thresholds of some adults, children, or animals may be lower. Of course, there's nothing special about lasers; any intense light source can be dangerous. Staring at the sun or an electric arc welder can also lead to eye damage. Responsible and safe handling of lasers is incumbent on all users; they aren't toys and should ***never*** be pointed at any person, animal, or vehicle.

The following is a list of surplus laser and laser equipment suppliers. It's not an exhaustive list, and inclusion or omission of any supplier does not imply endorsement or otherwise.

MKW Industries, 1440 S. State College Boulevard, Building 3B, Anaheim, California 92806. Telephone: (800)356-7714 or (714)956-8497. Laser surplus specialists. Lasers, power supplies, modulators, optic, books, and project kits.

Edmund Scientific, 101 E. Gloucester Pike, Barrington, New Jersey 08007. Telephone: (609)573-6259. New lasers, power supplies, and optics.

Marlin P. Jones, P.O. Box 12685, Lake Park, Florida 33403. Telephone (407)884-8764. Surplus electronics, some lasers power supplies, and optics.

Heathkit, P.O. Box 8589, Benton Harbor, Michigan 49022-8589. Telephone: (800)253-0570. "Laser Training System," 0.4 to 0.9-mW laser, power supply, and receiver. Laser can be amplitude modulated up to 10 percent.

Jerryco, 601 Linden Place, Evanston, Illinois 60202. Telephone (312)475-8440. Surplus electronics, optics, and miscellaneous items. Sometimes carries lasers and power supplies.

Those of you wishing to learn more about lasers might want to check out the following:

The Laser Cookbook, by Gordon McComb. TAB Books Inc., Blue Ridge Summit, Pennsylvania 17294-0850. Covers the basic principles of the laser, construction of laser power supplies, and a number of laser related projects (holography, light shows, fiberoptics, interferometry).

Understanding Lasers, Radio Shack 62-1333, $4.95. Good introductory guide to the principles and operation of lasers.

"Commmunicating on 474,038 GHz," by Steve Noll, WA6EJO, *Ham Radio*, December 1986, page 10.

Next month I'll talk about optical receivers, including their design and how to estimate sensitivity and "noise figure."

Microwave product news

Avantek has recently announced a component which may be of interest to those building microwave equipment. It's an active double balanced mixer/IF amplifier, type IAM-81018, and is priced at $38.90 (small quantities). The device is packaged in surface mount configuration (0.165 inch square) and consumes only 60 mW at 5 volts. For an RF (2 GHz) to IF (250 MHz) conversion gain of around 8 dB, only −5 dBm of local oscillator power is required. (From the spec sheet it looks as if operation with only −15 dBm LO power is possible with reduced conversion gain.) The IF output range is from DC to 1 GHz and the RF input range is from 50 MHz to 5 GHz. It appears that this device could be the heart of a multiband receive converter system, as it's a wideband device with 50-ohm input and output matching. The effective SSB noise figure at 2 GHz is 15 dB, so preamps would be required for good performance.

Well, that's all for now. Please send any questions, comments, or column ideas to me at 103 Division Avenue, Millington, New Jersey 07946.

✓ 109

BUILD YOUR OWN SUPERCHARGER

Charge 75 NiCds at one time

By W. C. Cloninger, Jr., K3OF, 4409 Buckthorn Court, Rockville, Maryland 20853

Handheld transceiver (HT), cordless screwdriver, rechargeable flashlight, extra HT battery pack, portable frequency counter, tape recorder... Do these items sound familiar? Most of them, and many other electrical devices, have something in common — rechargeable NiCd (nickel cadmium) batteries.

Most rechargeable devices come with their own charger, usually a small AC or DC wall transformer. They are generally single purpose. They charge a discharged battery to full charge, typically in 10 to 16 hours. They aren't designed for continued or trickle charging and may damage the battery if left on after the normal full charge period.

It's often desirable to keep NiCds completely charged and ready for use. Some of your rechargeable devices may sit for months before you use them. The solution to continual readiness is to keep your batteries on a maintenance or trickle charge.

Here's how to charge and trickle charge with the same unit. And why limit yourself to one charger? You can build a NiCd charger that will maintain the batteries in all of your rechargeable devices at the same time.

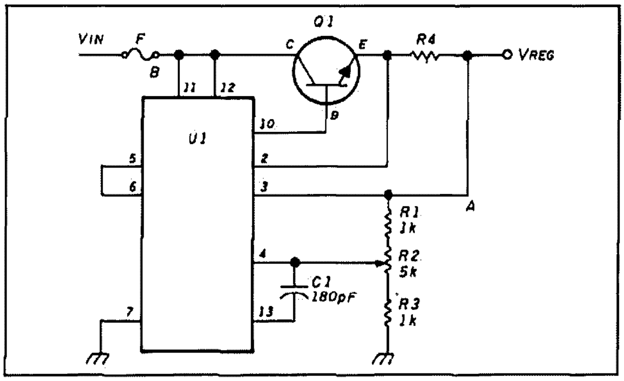
I've built a device that I call the Supercharger. The one shown in Photos A and B will charge and/or trickle 75 or more NiCds spread among up to six different rechargeable devices!

Photos C through E show the Supercharger in various stages of assembly to give you an idea of parts placement and wiring arrangement.

Theory of operation

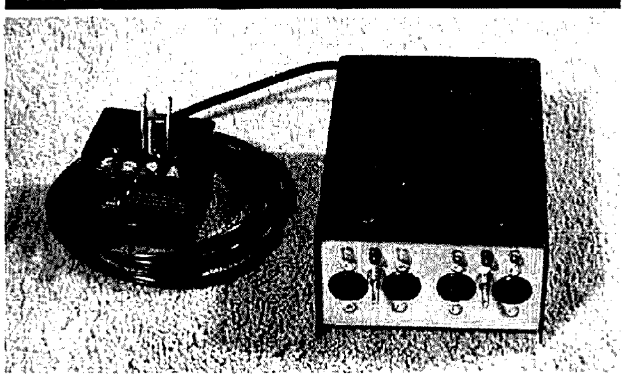
The basic building block of the Supercharger is a simple DC power supply with current limiting and adjustable voltage. The circuit shown in Figure 1 uses the popular 723 voltage regulator. This regulator was discussed in a previous *Ham Radio* article.[1] The Supercharger uses fewer components than a regular power supply because it doesn't need RF protection. All components are inexpensive and readily available.

FIGURE 1



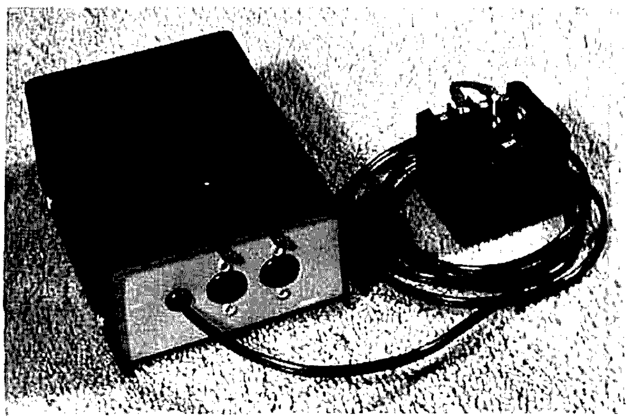
The basic 723 regulator circuit. Vin is typically 15 to 28 volts to suit your needs.

PHOTO A



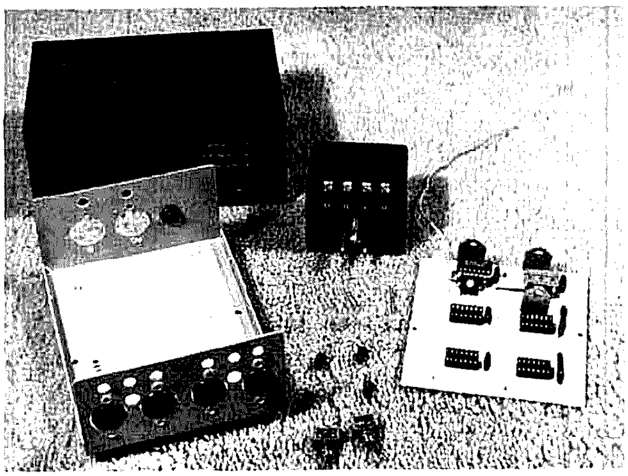
The Supercharger with four of the six outputs located on the front panel. The switches are for high/low charge rates on two of the outputs.

PHOTO B



Rear view of the Supercharger showing two output sockets on the rear panel.

PHOTO C



Major Supercharger components and partially completed pc board.

Current limiting of the device is determined by resistor R4. Use the following formula:

$$Current\ limiting = \frac{0.65}{R} \quad (1)$$

A 6.5-ohm resistor for R4 will limit current to 100 mA. R2 is for voltage adjustment and is normally set to a level higher than the maximum voltage required by the device you are charging. Of course the input voltage to the Supercharger must be sufficiently high.

The most popular NiCds are AA cells which are usually 450 mAH (milliampere hours), and C or D cells which are usually 1.2 AH (ampere hours). The normal charge rate lor AA cells is 45 to 50 mA for 14 to 16 hours (150 mA for 4 hours for quick charge cells). The usual charge rate for C or D cells is 80 to 100 mA for 14 to 16 hours.

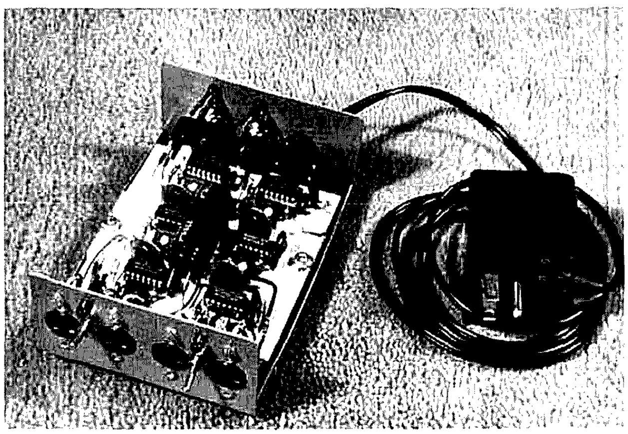
Design criteria

The Supercharger is designed to provide charge rates of approximately 50 mA, 100 mA, and 150 mA. Lower trickle charge rates of 15 mA are provided for AA cells and 30 mA for C and D cells. **Figure 2** shows the R4 values needed to provide three charge levels and two trickle levels. **Figure 2A** shows a single current output circuit. **Figure 2B** shows how R4 may be switched to provide high/low (charge/trickle) charge rates. I added LEDs because I like to "see" what's happening. For 15 mA and 30 mA rates, you can use LEDs directly without any current-sharing resistors. They will light only when current is actually flowing. The low charge LED in **Figure 2B** is really an indicator of switch position, but it doubles as a power indicator for the Supercharger. The LED across R5 lights at high charge levels only when the voltage drop across R5 is great enough for current to flow through this LED. The switches and LEDs were added strictly for personal preference and aren't required.

Circuit board

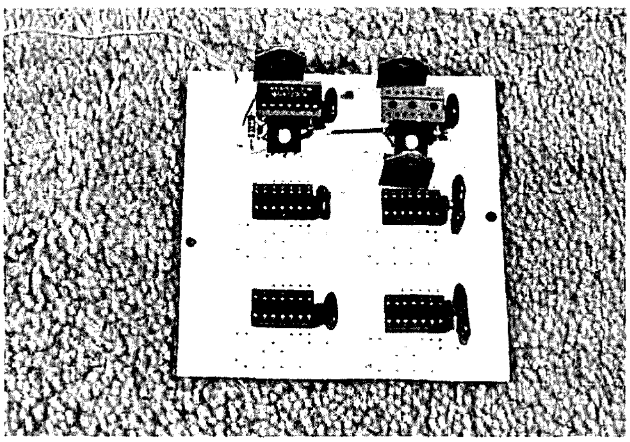
The circuit board for the Supercharger is shown in **Figure 3** for eight different charger outputs. My unit uses only

PHOTO D



Completed Supercharger with cover removed. Two of the outputs are located on the rear panel. Note the heat sinks on the two pass transistors at the rear of the pc board.

PHOTO E



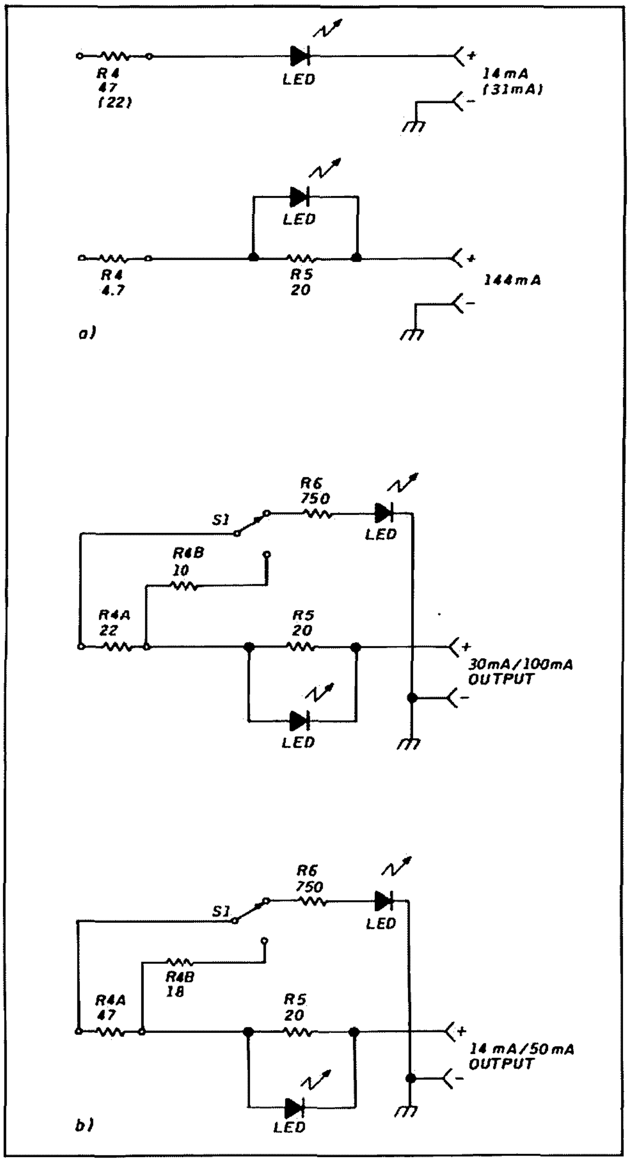
This Supercharger uses six simple regulator circuits on the same pc board. The black wire at the rear is a jumper to connect the negative supply to both halves of the board.

six outputs because that's what my minibox would accommodate. It's easy to make the pc board using the TEC-200™ film method.[2]* It took about an hour — not counting a trip to a local copy shop.

Construction

Once you've made the board, finish assembling the unit and mount it in a suitable enclosure. Component placement for a single charger output is shown in **Figure 4**. Q1 can be almost any NPN power transistor that will handle the load. The pin configuration of Q1 may vary, but most transistors can be mounted directly on the pc board with proper positioning or perhaps turning them

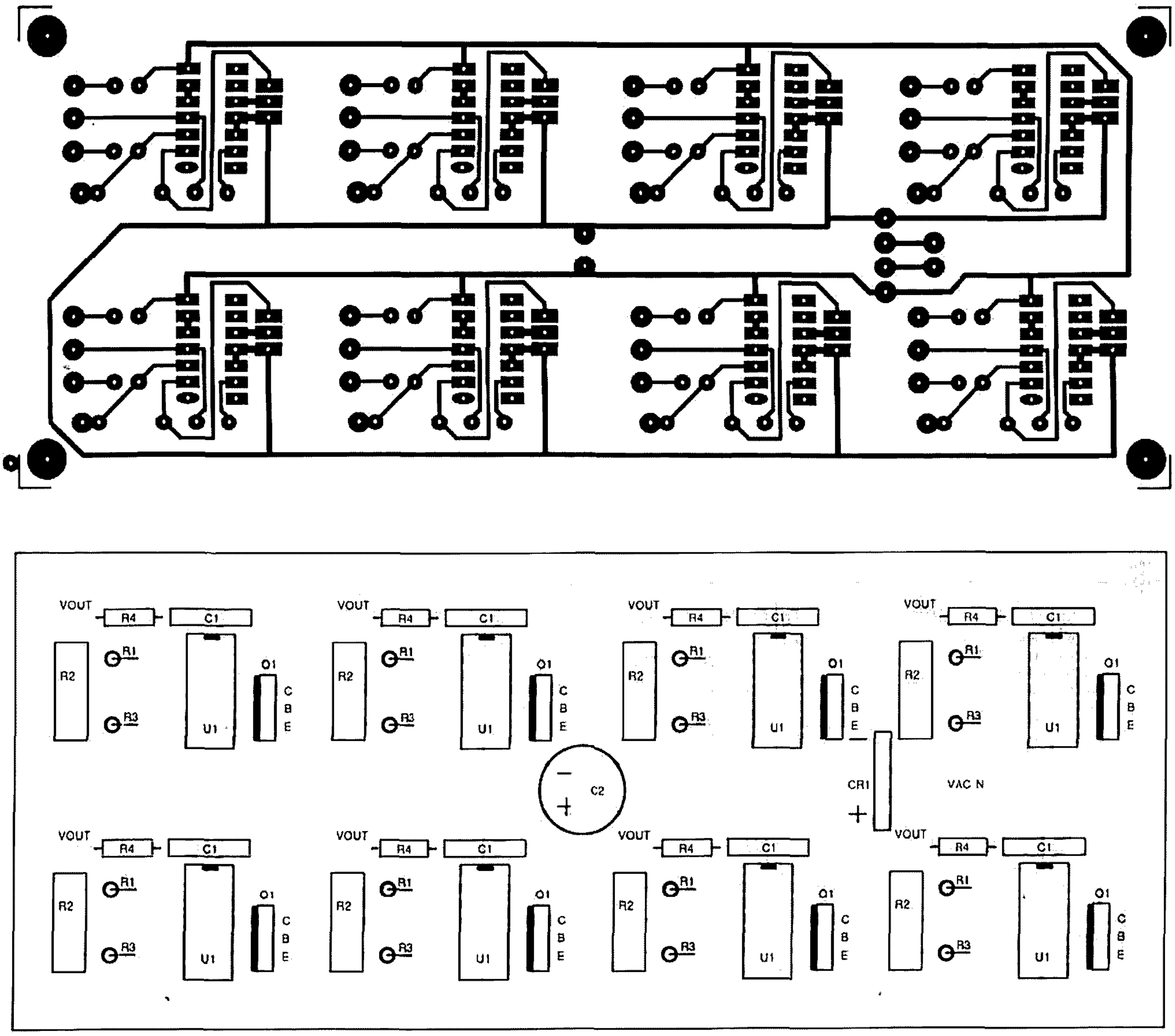
FIGURE 2



Single current output, (A), LED and/or R5 are optional. High/low selectable output circuit, (B).

* The Meadowlake Corporation, 25 Blanchard Drive, PO Box 497, Northport, New York, 11768. (516)757-3385

FIGURE 3



Foil and component sides of the pc board with eight regulator circuits. The board can be cut for the number of circuits desired. CR1 and C2 are mounted on the component side. CR1 may be omitted if Vin is DC.

180°. On the TIP35 I used, I bent the emitter leg 90°, let it lie flat on the top of the board, and soldered the jumper directly to the leg. You'll have several unused pads on the pc board because it includes the capability for an RF-protected power supply.

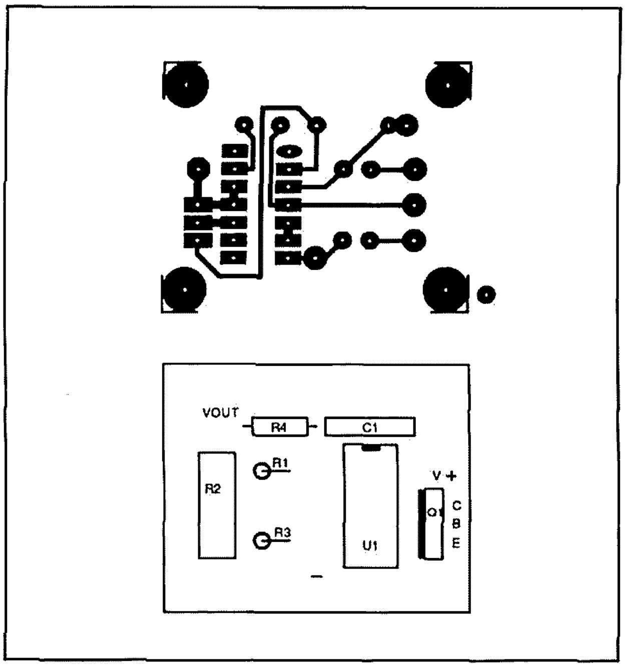
Input power to the Supercharger may be AC or DC. I used a 22-volt 0.6-A AC wall transformer, so I added full-wave bridge CR1 and filter capacitor C2 directly to the board. Some wall transformers already supply DC voltage and can be used without CR1 by connecting the DC leads to the appropriate busbars on the pc board. By the way, if you install CR1, you can still use your DC supply without regard to input polarity. You must also place jumpers from positive to positive and negative to negative on each "bank" of four circuits.

Operation

Now power up the unit. My DC supply was about 28 volts (less under load), so I adjusted R2 for 20 volts for all six outputs. I checked the current limit of each regulator by placing an ammeter directly across the output.

Each output will charge 1 to 12 NiCds easily without any adjustment. Remember, all you care about is the proper current, and that's determined by R4. If you are charging only one NiCd, there will be a large voltage drop across Q1. This power dissipation may require a heat sink on Q1. If you use a full charge rate, don't forget to change to trickle charge after the appropriate charge time by adjusting the switch position or plugging your device into an output of lower current.

FIGURE 4



Foil and component sides of pc board with one regulator circuit. The board allows some flexibility for mounting Q1 and R2. R1 and R3 are mounted vertically.

Automatic fast charge — a bonus

There's one more neat trick you can do with the Supercharger. Because R2 was included and will adjust voltage from about 8.5 to 25 volts, you can set the voltage limit to reduce the charge rate automatically as your NiCds reach full charge.

To determine the proper voltage limit for a particular battery pack (my HT battery, for example), I give it a normal full charge using the charger supplied by the manufacturer. Then I place the battery pack on one of the 150-mA outputs with an ammeter in series. Next, I adjust the voltage using R2, so that the current to the fully charged battery pack is 15 mA (0.03 × 500 mAH, or 15 mA for an AA cell).

If I place a discharged HT battery pack on this output, it charges immediately at 150 mA. The charge rate tapers off as the battery becomes charged. This "fast charge" isn't as efficient as a sophisticated circuit using comparators and/or timers, but it sure beats the wall charger that comes with the HT. And, I didn't have to do anything to the Supercharger to get this added benefit.

REFERENCES

1. W.C. Cloninger, Jr., K3OF, "Understanding and Using 723 Voltage Regulators," *Ham Radio*, March 1989, page 42.
2. W.C. Cloninger, Jr., K3OF, "EZ PCBs," *73*, August 1987, page 43.

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1. Scott Hoter, N7DFR, "Tech Tips," *73*, June 1988, page 80.
2. Dr. Roland Milker, DL2OM, "Technical Correspondence," *QST*, May 1988, page 42.

EFFECTIVE NOISE TEMPERATURE PART 2

EQUIVALENT NOISE TEMPERATURE AND NF

Understanding optimum noise performance

By Michael E. Gruchalla, P.E., 4816 Palo Duro N.E., Albuquerque, New Mexico 87110

In "Effective Noise Temperature, Part 1," I gave an introduction to and background on the concept of effective noise temperature. Part 2 picks up the discussion of the limiting value of source voltage, and goes on to talk about noise figure phenomena and amplifier equivalent noise temperature.

When the source in the small signal example in part 1 is fed to a matched 50-ohm load, the load signal and noise voltages are each 0.561 μV. The power delivered to the load is 6.29×10^{-15} watts. That is precisely the power you'd expect to be available to a matched load from a room temperature resistive source and π/2-MHz noise bandwidth using **Equation 1** from part 1.

Noise power and a matched load

It's important to note that the noise power available to a "matched" load is independent of the source resistance. This is implied by **Equation 1**. It's a bit difficult to see, because *if you look at* **Equation 5** *from part 1 you'll note* that changing the source resistance changes the equivalent noise voltage (or current in **Equation 6**, part 1). However, if you change the source resistance — lower it to reduce thermal noise voltage, for example — the matched load is changed to the same value. You can take the model of **Figure 3**, part 1 in terms of the general source resistance and compute the noise power delivered to the load using **Equation 5**.

$$\begin{aligned} P_L &= E_L^2/R_L \\ &= [(R_L/(R_L+R_s))\ (2\sqrt{k\ T\ BW_n\ R_s})]^2/R_L \\ &= [R_L/(R_L+R_s)]^2\ 4\ k\ T\ BW_n\ (R_s/R_L) \end{aligned}$$

but $R_L = R_S$ for a matched load. Thus:

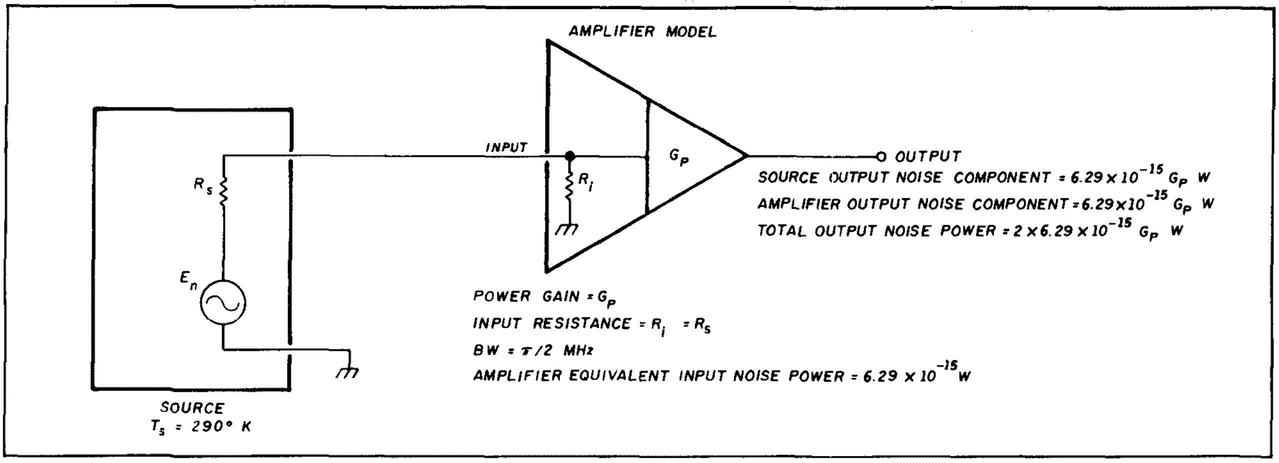
$$P_L = k\ T\ BW_n \qquad \textbf{(7)}$$

The result in **Equation 7** is the same as that given in **Equation 1**. This is to be expected because you worked forward from **Equation 1** to arrive at **Equation 5**. However, taking the analysis full circle to the starting point should demonstrate that the thermal noise power available at a matched load is independent of the actual resistance, even though the actual equivalent noise voltage and current are functions of the resistance.

Maximum thermal noise power

The implication of **Equation 1** is very important. Because the noise power available at a matched load is independent of the source impedance (resistance and reactance), you may compute the maximum thermal noise power available to a load (an amplifier input, for example) without any information about the amplifier or the system in which it is used. From **Equation 1**, that power level at room temperature is about 4×10^{-21} watts/Hz of noise bandwidth, or −174 dBm/Hz. This may seem like a very small level, but remember the value is for a "1-Hz" bandwidth. In the preceding example with a π/2-MHz noise bandwidth, the limiting thermal noise power would be −112 dBm. The 1.12-μV thermal noise voltage of the 50-ohm source resistor with a π/2-MHz noise bandwidth delivered 0.56 μV to a matched load. Consequently, the thermal noise power available at the load from the source resistance is 6.29×10^{-15} watts, or −112 dBm.

FIGURE 1



Amplifier signal-to-noise model.

Source noise and amplifier noise

It may seem that the noise performance (S/N) could be improved by providing a less than optimum impedance match. It's true that in such a case less source noise would be available at the amplifier input. However, because the source resistance is the same for both the noise and signal components, less signal will also be coupled. For any given signal, the source signal-to-noise ratio remains constant for any matching condition. Because the amplifier to which this source is attached is very likely to contribute some noise, any reduction in the input signal component tends to cause the amplifier noise to be more significant and suggests a degradation in the noise performance of the system; that is, the output signal-to-noise ratio would be degraded. But for any specific amplifier and nonzero source level, there's a source impedance that will result in the best signal-to-noise ratio. This is the noise match impedance, and it's totally unrelated to the optimum power transfer impedance.[1,3] It is a function only of the amplifier noise characteristics. This concept is discussed in detail in **References 1 and 3.**

Noise figure

The noise figure is defined as the ratio of the total output noise power of a system to the output noise power due to the source alone, expressed in decibels. You can read **Reference 3** if you'd like a detailed description.

$$NF = 10 \log\left[\frac{\textit{Total Output Noise Power}}{\textit{Output Noise Power Due to the Source}}\right] \quad (8)$$

Noise analyses are generally referred to the output of the systems under consideration. The system output is typically the most convenient point to make signal measurements. Also, in many cases, the various signal quantities of interest aren't directly accessible. For example, the equivalent input noise voltage of an amplifier can't be measured directly. It must be computed from output noise measurements.

Consider the amplifier in **Figure 1** with a power gain Gp, signal bandwidth of 1 MHz, equal source and input resistances, and a room temperature source. The noise power at the amplifier input due to the source is 6.29×10^{-15} watts. Also, let the amplifier have an "equivalent input noise power" equal to that delivered to the input by the source resistance, 6.29×10^{-15} watts (unity power signal-to-noise ratio). The output noise power due to the source is 6.29×10^{-15} watts $\times$ Gp. The total output noise power is equal to the sum of the source and amplifier noise power contributions. Because the amplifier equivalent input noise was defined to be equal to the source noise, the total output noise power is twice the source noise contribution. This means the noise figure is 10 log (2), or 3 dB.

Noise figure as a function of temperature

The preceding seems like a reasonably good specification for the noise performance of an amplifier. It is perhaps the most common method of specifying the noise of various systems. However, it does have at least one serious shortcoming: a specific noise figure value is valid only at a *specific temperature.* For a specified noise figure value to apply, the source must be at the same temperature in the application as it was for the original specification. To demonstrate this, I'll cool the source resistor in **Figure 1** to a temperature of 96.7°K. The noise figure given by **Equation 8** will then be 6 dB. The cooler resistor contributes less noise. So the amplifier noise power, although the same in both cases, is a factor of 3 higher than the source contribution when the source is at 96.7°K, while it is equal in the 290°K case. The total output noise is actually reduced with the colder source. For any given signal, the signal-to-noise ratio is increased. But the percentage of total output noise contributed by the amplifier in the cold case is higher than it is for the room temperature case. This results in a poorer noise figure, even though the signal-to-noise ratio for any specific signal is improved (see **Reference 3** for a more detailed discussion of NF versus S/N). In order to effectively use a noise figure specification, you must know the temperature at which the noise figure was specified. Of course, this isn't particularly useful if you wish to use the amplifier for sources of a different temperature, or perhaps if you don't precisely know the source temperature.

Equivalent noise temperature

The noise characteristics of an amplifier may be specified in an alternative manner. From **Equation 1**, the maximum noise power available at the input to a system is a function of only the source temperature and the noise bandwidth (and of course Boltzmann's constant). This maximum available noise power occurs only for a power-matched condition of the source and amplifier input. For any real system, a value of temperature for the matched source may be computed by rearranging **Equation 1** so the noise contribution due to the source at some point in the system of interest (usually the output as discussed above) is the same as that contribution at that same point from the noisy amplifier itself.

> The equivalent noise temperature of the amplifier is the temperature to which a power-matched source resistance must be set to provide a source output noise power component equal to the amplifier component of output noise power.

In the amplifier example with the room temperature (290°K) source, the source noise contribution and amplifier noise contribution are equal. Here, the source must be at 290°K to provide an output noise component equal to that of the amplifier. That amplifier is then a 290°K amplifier. If this amplifier were used with the 96.7°K source, it would still be a 290°K amplifier. It's important to remember that equivalent noise temperature is a property of the amplifier (or other system characterized) — somewhat like gain — and is independent of the application.

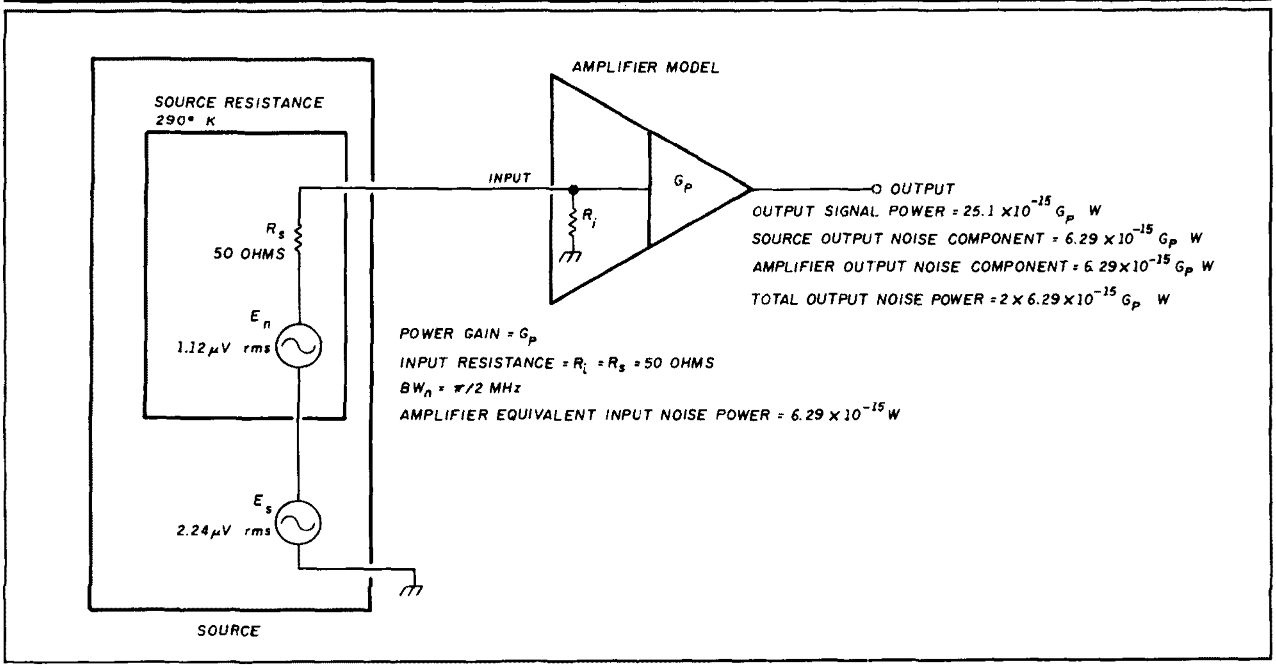
Ideal and real noise applications

The noise temperature specification has a shortcoming. It is a specification of noise performance when the source is power matched to the amplifier and may be thought of as a type of ideal specification. It usually doesn't represent the performance provided in a real application. If an unmatched source is used with an amplifier of some specified noise temperature, the output signal-to-noise ratio will be different from that obtained with ideal matching (it could be better or poorer). However, the noise temperature specified for the amplifier would still be correct because it's based on performance in a specific configuration. Further, because the noise temperature is based on an input power-matched condition, it provides no information about the optimum performance that may be obtained in a noise-matched input. A similar noise temperature versus signal-to-noise ratio comparison could be made as presented in **Reference 3** for noise figure versus signal-to-noise ratio. Just because an amplifier has a very low noise temperature doesn't mean it will provide best signal-to-noise ratio in a specific application. It's very likely that a higher noise temperature amplifier with an input well noise matched[3] to the source will provide a much better signal-to-noise ratio than an amplifier of much lower noise temperature with a poorly noise-matched input. The optimum noise-matched source resistance is given by the ratio of the equivalent input noise voltage and the equivalent input noise current. See **References 1 and 3** for a thorough review of noise matching.

Noise-matching example

• **290°K** amplifier. Consider the amplifier in **Figure 2**. It's the same as that of **Figure 1** with a signal added. The amplifier equivalent noise temperature is 290°K, input resistance is 50 ohms, signal bandwidth is 1 MHz, and power gain is Gp. For convenience, I chose 2×1.12 μV rms as the input source signal. This is twice the source thermal noise level. The signal at the amplifier input is then 1.12 μV rms and the output

FIGURE 2



290°K amplifier model.

put signal power 25.1 × 10^{-15} Gp watts. The noise voltage available at the amplifier input from the 290°K source is 0.561 μV, an output noise power due to the source of 6.29 × 10^{-15} watts. Because the amplifier has a 290°K noise temperature, its equivalent input noise power is the same as that of the 290°K source, or 6.29 × 10^{-15} watts. As a result, the total output noise power is 12.58 × 10^{-15} Gp watts. This means the power signal-to-noise ratio is 2, or 3 dB.

• **145°K** amplifier. With this same source, I'll try another amplifier with a noise temperature of 145°K. This is a factor of 2 lower than the 290°K unit, but with a 10-ohm input resistance — a factor of 5 lower than the matched case. I chose these values for convenience of calculation; they aren't intended to represent any realistic application. However, the 145° value is a modest specification, and the 10-ohm value of input resistance isn't too unreasonable. An NE13783 low noise FET exhibits about a 10-ohm resistive input impedance at 12 GHz. These values are actually quite realistic.

This amplifier is shown in **Figure 3**. Because this is a mismatched case, you need additional information about the amplifier to make truly accurate computations. The noise voltage and current sources at the amplifier input are the "equivalent input noise sources."[1] The values shown give an optimum noise match and 145°K noise temperature at the specified 10-ohm source resistance. There's a family of equivalent input voltage and current source values that will provide a 145°K noise temperature. In a real application of computing noise performance, the equivalent sources would be accurately measured. For this example, the values of **Figure 3** are the defined amplifier characteristics.

• **Noise temperature verification**. First I'll verify that the amplifier is a 145°K unit. The input noise power due to the two amplifier noise sources alone when attached to a matched 10-ohm noiseless (zero degree) source is 3.14 × 10^{-15} watts, and the output noise power is 3.14 × 10^{-15} Gp watts. According to **Equation 1**, the temperature to which a souce resistance must be raised to deliver that thermal noise power to a matched load with a π/2-MHz noise bandwidth is 145°K. The amplifier is then indeed a 145°K unit.

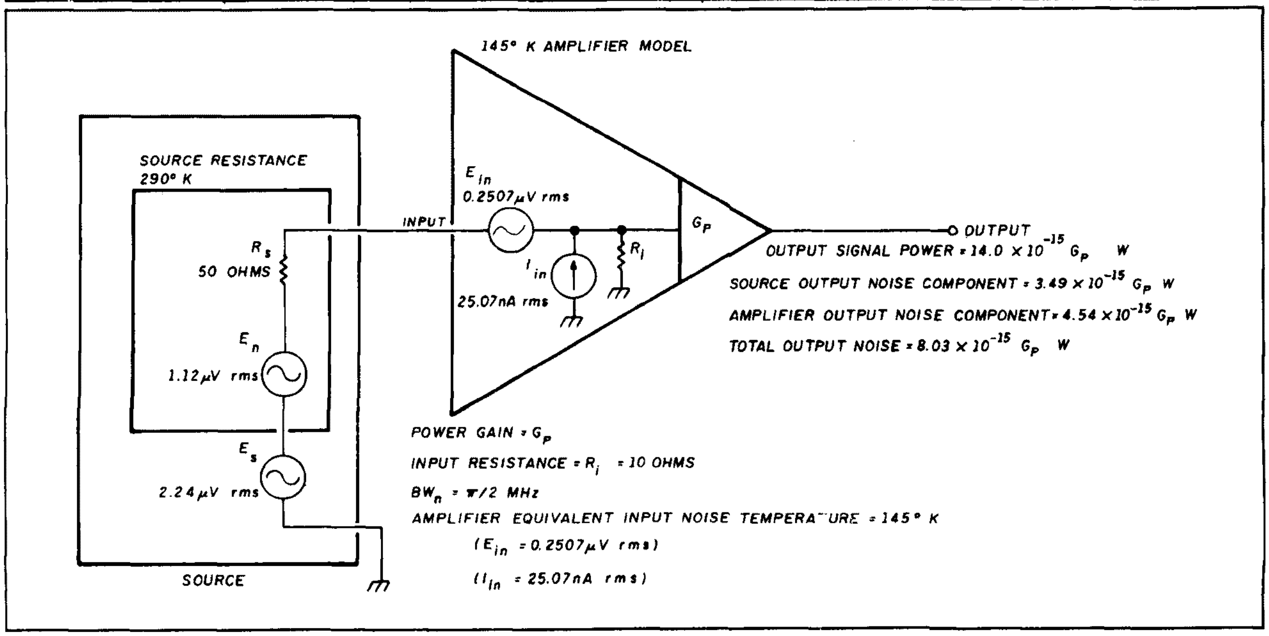
The noise power at the amplifier input due to the 290°K, 50-ohm source and the amplifier equivalent noise sources in **Figure 3** is 8.03 × 10^{-15} watts. This yields a total output noise power of 8.03 × 10^{-15} Gp watts. The input signal voltage is 0.374 μV which produces an output signal power of 14.0 × 10^{-15} Gp watts. The power signal-to-noise ratio is then 1.74, or 2.4 dB.

• **Analysis**. Substituting a lower noise amplifier in the system with the 50-ohm source resulted in a poorer signal-to-noise ratio. But what would happen if you had a source with a 10-ohm source resistance and the 145°K amplifier? This is the optimum "noise-matched" resistance for the amplifier in **Figure 3**. (It's also the optimum power match in this case, but that isn't significant.[3]) This impedance transformation may be very conveniently and realistically accomplished using a transformer with a 2.24:1 turns ratio. By using an ideal transformer, you maintain the source signal power constant and can easily compare various power signal-to-noise ratio results. The transformed source temperature remains 290°K. The amplifier noise temperature is still 145°K because that specification is independent of application. In this case the input signal power will be 25.1 × 10^{-15} watts and the output signal 25.1 × 10^{-15} Gp watts. The total output noise power from the 10-ohm, 290°K source and the amplifier noise sources will be 9.43 × 10^{-15} Gp watts. That results in an output power signal-to-noise ratio of 2.66:1, or 4.3 dB.

All noise parameters must be recognized

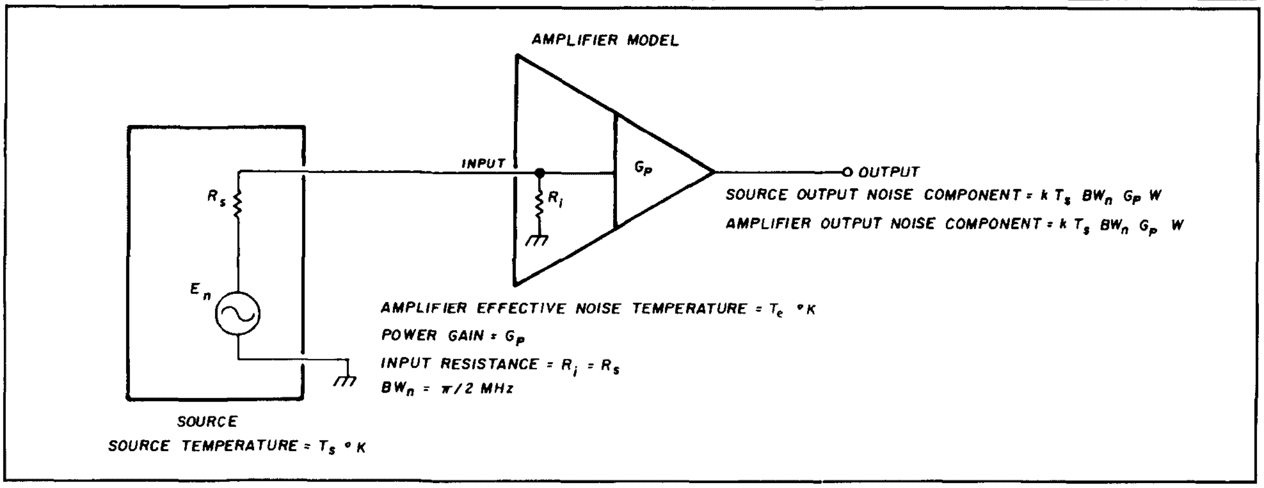
I started with a 290°K amplifier with a 50-ohm input resistance and a 290°K, 50-ohm source and obtained a 3-dB signal-to-noise ratio with a 2.24-μV source signal voltage.

FIGURE 3



145°K amplifier model.

FIGURE 4



Model for comparing noise figure and effective noise temperature.

I then substituted a lower noise 145°K amplifier with the same signal source and obtained a 2.4-dB signal-to-noise ratio. The lower temperature and lower noise amplifier provided a 0.6 dB poorer signal-to-noise ratio than the higher noise 290°K amplifier with the 50-ohm room temperature source. Then I transformed the 50-ohm, 290°K source to a 10-ohm optimum noise-matched value, still at 290 °K. With the same 145°K amplifier that provided the 2.4-dB signal-to-noise ratio with the 50-ohm source, I then obtained a 4.3-dB signal-to-noise ratio. So this 145°K amplifier with a fixed source power provided a 2.4-dB signal-to-noise ratio with one source resistance and a 4.3-dB value with another. But the noise temperature was constant. Can this be correct? Is there an error here?

The calculations are correct, and indeed this result can easily occur in practice. Although the higher noise temperature amplifier generally exhibits higher noise than the lower temperature unit, it provides a better noise match to the 50-ohm source than did the specified 145°K unit. This gives a better signal-to-noise ratio for any given signal. In the final example, the 10-ohm source provides an optimum noise match resulting in the best signal-to-noise obtainable with the particular amplifier being modeled. What's important here is that the lower noise temperature amplifier has the *capability* to provide the better noise performance. But to actually obtain improved performance, all of the noise processes must be well understood and the various noise parameters used correctly.

Importance of understanding noise parameters

In practice you rarely know the exact resistances involved. You have to ask: What is the actual source resistance of an antenna? Is it the radiation resistance? The element resistance? The connecting transmission line resistance? Or is it a combination of all of these? What is the effective temperature of the source(s)? Further, what is the actual input resistance of a low noise amplifier (LNA)? Worse yet, does the LNA even come close to providing a good noise match to your source? These are all very difficult questions to answer. Careful measurement of the noise parameters must be completed before you can accomplish any true optimization. It's actually possible to replace one LNA with a lower noise temperature unit and achieve a poorer signal-to-noise ratio. Those of you who have experimented with various "inexpensive" LNAs may have stumbled on this paradox. Because these amplifiers are manufactured more for low cost than consistency, they tend to exhibit considerable variation in performance parameters from unit to unit. Even when substituting two identical models, you may see a considerable difference in noise. Of course, a noisier unit may also be defective, or a quieter one could be a premium low noise unit. An understanding of the actual noise processes is critical for successfully optimizing system noise performance.

Comparing noise temperature and NF

As a final thought, it might be useful to compare noise temperature and noise figure. Both of these specifications are measures of device noise, so it should be possible to compare them effectively. To do this, use **Equation 8** and the amplifier model in **Figure 4**. This amplifier configuration is similar to those shown earlier, but has a general source temperature T_S and an amplifier equivalent temperature T_e. Because the configuration in **Figure 4** offers a matched load to the source, you know the noise power delivered to the amplifier input from **Equation 1**. The output noise power due to the source alone, P_S, is then given by **Equation 9**.

$$\textit{source output noise power} = P_s = k\ T_s\ BW_n\ Gp \qquad \textbf{(9)}$$

Now when the amplifier is attached to a power-matched source at a temperature equal to the effective temperature of the amplifier, the source and amplifier contribute equal noise components to the total output noise. This is essentially the definition of effective noise temperature. The component of the amplifier output noise, P_a, is simply equal to the noise that would be output if the amplifier were noiseless and the source were at a temperature T_e. By the definition of effective temperature, T_e is that temperature to

which you must raise the source if the amplifier were noiseless to produce an output noise equal to that of the actual noisy amplifier with a noiseless source. Once again, the input power is given by Equation 1 and the corresponding amplifier output noise power contribution by Equation 10.

$$\textit{amplifier output noise power} = k\, T_e\, BW_n\, Gp \qquad \textbf{(10)}$$

By combining Equations 8, 9, and 10, you can express the noise figure in terms of the source temperature and the amplifier effective temperature.

$$NF = 10 \log \left[\frac{P_s + P_a}{P_s}\right]$$

$$= 10 \log \left[\frac{(k\, T_s\, BW_n\, G_p)\ (k\, T_e\, BW_n\, G_p)}{(k\, T_s\, BW_n\, G_p)}\right]$$

$$= 10 \log \frac{T_s + T_e}{T_s}$$

$$NF = 10\ Log\ [1 + T_e/T_s] \qquad \textbf{(11)}$$

Equation 11 then lets you compute the noise figure of an amplifier from its effective noise temperature. For example, if you have a room temperature source of 290°K (T_s = 290°K), a 290°K amplifier (T_e = 290°K) will exhibit a 3-dB noise figure as expected. You may also compute the effective noise temperature from the noise figure by solving Equation 11 for T_e. That result is shown in Equation 12.

$$T_e = T_s\ (10^{[NF/10]} - 1) \qquad \textbf{(12)}$$

Closing remarks

Hopefully, all this has shown you just what the concept of effective noise temperature means and how it was derived. Unlike noise figure, the noise temperature is a device parameter (like gain) and is independent of the application of the device. On the other hand, noise figure is a comparison of the output noise power of a system due to the source alone — a real source, any source, matched or not — to the total output noise power of the system. So, the noise figure is a type of "practical" noise specification showing the performance of a real system, while the noise temperature is a type of standard specification defined under very specific conditions.

Both equivalent noise temperature and noise figure are useful parameters, but both have shortcomings. The successful use of any parameter depends upon a thorough understanding of that parameter. Obtaining optimum noise performance depends greatly on your understanding of the noise processes and careful application of the optimizing principles discussed in this article and the references.

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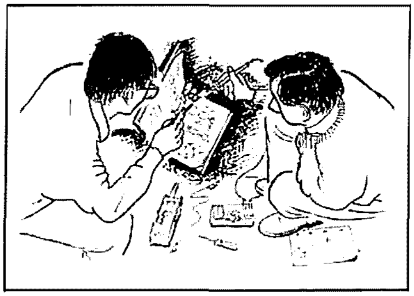


An RF Current Loop

Here's a handy device for tracking down RF current in your antenna. Knowing your antenna means knowing where RF current flows. Does your feedline radiate? It shouldn't, even if it's made of open wire line. How about the phone lines or guy wires? These and other objects could be robbing you of RF or distorting your radiation pattern. The circuit shown in **Figure 1** and pictured in **Photo A** will give you a true representation of your antenna by detecting RF current.

PHOTO A

Completed RF tracker.

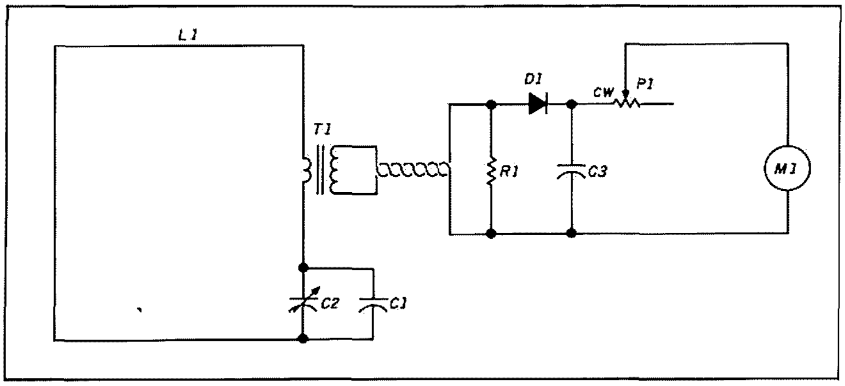
TABLE 1

Loop components versus frequency of operation.

Frequency, MHz	*Wire loop L1, inches	C1, pF	C2, pF
16 - 30	15	none	95 - 420
9 - 16	26	none	95 - 420
7 - 8	26	560	95 - 420
3.7 - 4.0	26	2200	95 - 420

* Cut length ol copper-weld wire

FIGURE 1



RF current loop.

PARTS LIST

C1	*Silver mica. See* Table 1 *for value.*
C2	*95 to 420-pF trimmer (Radio Shack)*
C3	*Ceramic capacitor, 0.1-μF, 100 volt*
CR1	*Diode, 1N914*
L1	*Wire loop. See* Table 1 *for length.*
M1	*Meter, 0 to 1 mA*
R1	*Resistor, 56-ohm carbon composition 5 percent, 1/4 watt*
R2	*Potentiometer, 10 k*
T1	*T-50-2 (Amidon), wound with 15 turns no. 26 AWG — leave a 7" pigtail*

Limited range of C2 and tolerance of parts may require up to a 10-percent change of C1 value.

*** Amidon Associates, 12033 Otsego Street, North Hollywood, California 91607**

Operation

Load the antenna with about 50 watts. Run the loop along the length of the suspected object while observing meter deflection. Maximum meter indication occurs when the wire edge of the loop is parallel with the direction of current. A meter zero (null) will occur at angles perpendicular to current flow.

Remember that current peaks repeat every half wavelength, so that maximum meter indication may be one-quarter wavelength from where you initially probed. There will be little or no indication at voltage maxima.

Calibration

Carefully tune the loop using a grid-dip meter. Adjust C2 until resonance is indicated. You may want to monitor the grid-dip frequency with your station receiver to verify meter accuracy.

An alternative tuning method would be to hold the loop near the antenna and adjust C2 for maximum. The peak will be sharp, so adjust carefully.

The key to the loop's performance is its relatively high Q afforded by toroidal transformer T1. The loop should be operated within about 150 kHz of resonance for best sensitivity and results.

Construction

Shape a piece of no. 14 copper-weld wire into a square and thread it

through transformer T1. See Table 1 for loop length. Secure the loop to one end of a 5/16 by 24-inch dowel. Be sure to leave about a 1-inch gap in the loop to mount C2. Twist the leads of T1 and run them down the dowel about 6 inches.

Mount the detector components consisting of R1, CR1, and C3 to the dowel and wrap them with electrical tape. Attach the wires from C3 to the meter and sensitivity pot located near the handle.

Final remarks

I have discussed a simple means of detecting antenna currents with the aid of the RF current loop. Now that the currents can be identified, they can be contained. Past *Ham Radio* articles and the *ARRL Antenna Book* contain some excellent suggestions to help reduce unwanted radiation and pickup. It should now be easy to evaluate the best solution for your antenna system. Let's go for top performance and put that RF up where it belongs!

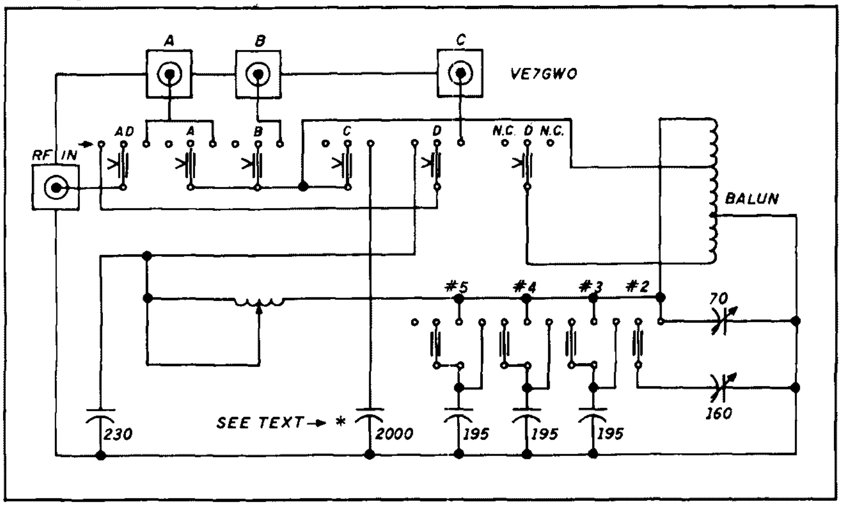
Tom Rehm, K9PIQ

Updating The Viking MB-VA Antenna Tuner

After purchasing and connecting a Viking antenna tuner MB-VA, I was disappointed because it would not tune my 160-meter antenna. Here are two solutions to the problem.

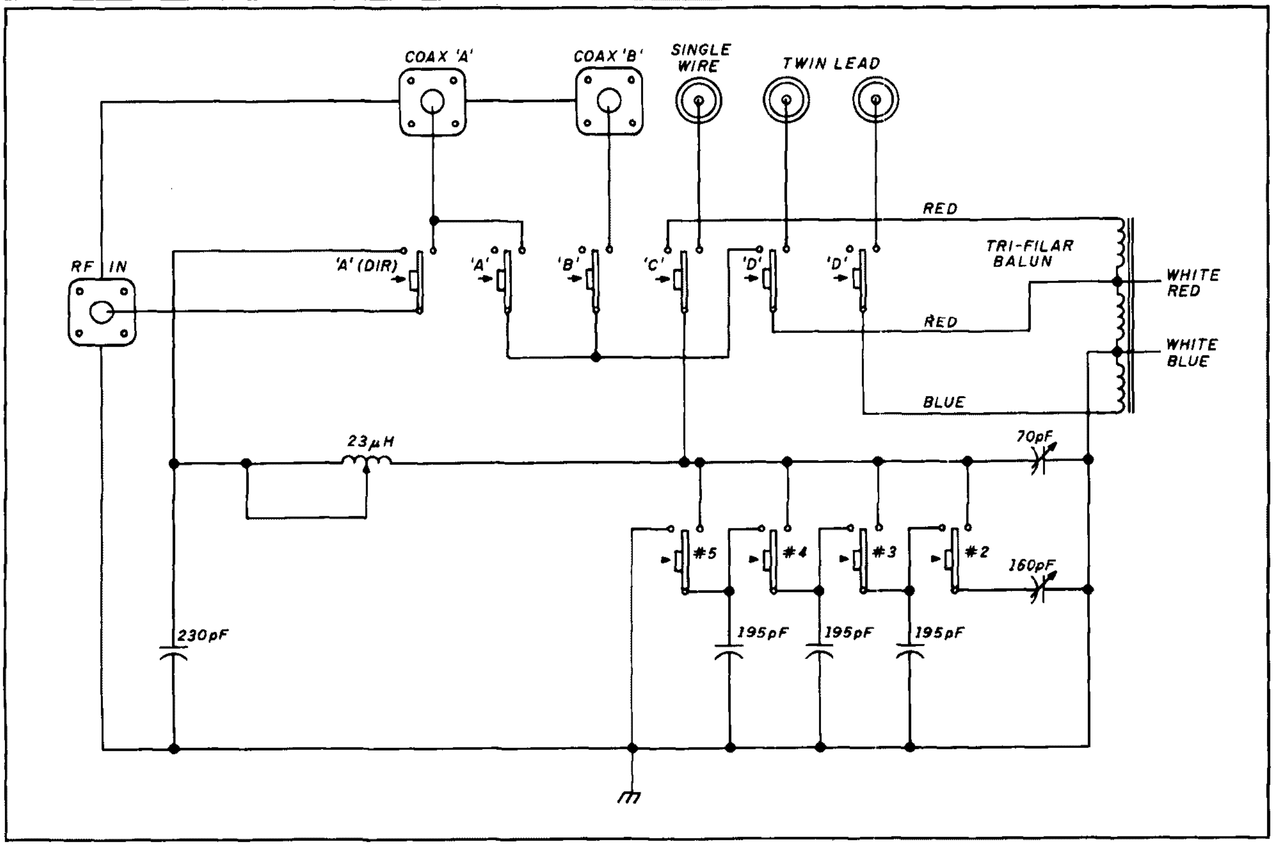
I could leave the antenna tuner wiring the way it was. Then, for 160-meter operation, I could mount a 2000-pF

FIGURE 1



Schematic showing changes to the Viking MB-VA antenna tuner.

FIGURE 2



Manufacturer's diagram of the Viking MB-VA tuner.

10-kV capacitor (on the skirt behind the roller inductor) from coax connector B through a high voltage switch to ground.

I didn't need a single wire or a balanced line input. The solution I chose involved the more elaborate changes shown in Figure 1. The modifications allow me to run my tribeam directly, or through a tuner on coax A and A direct. I can now tune 160 meters from the front switches, by pushing B and C *at the same time.* The switches release simultaneously when other switches are pushed.

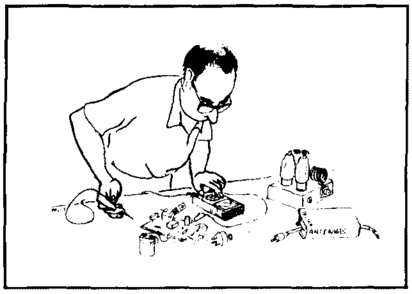
A further change lets me use a dummy load through connector C. I disengaged the connections from the insulators of the balance line, then removed the single insulator and replaced it with coax connector C.

Note that there is an error in the manufacturer's diagram. The bus connecting A and B *does not* connect to C (Figure 2).

I hope my suggestions will help Viking tuner owners make the most out of this well-made piece of ham gear.

G.W.T. Oliver, VE7GWO

Ham Radio Techniques



Bill Orr, W6SAI

160-METER ANTENNA PROBLEMS AND SOLUTIONS

The quarter-wave Marconi working against ground is a popular and inexpensive antenna for 160 meters. A lot of newcomers to the band favor this simple antenna because it's easy to put up, it isn't too big, and it works.

I erected such an antenna a few years ago. I had an enjoyable time and worked a lot of stations on 160, but the signal reports I received weren't very ego building. Worst of all, many of the local 160-meter crowd were working DX I couldn't even hear! That was a bad sign. It meant that something had to be done about the antenna. I couldn't go to a vertical, and the position of the house on the lot precluded putting more wire up in the air. I had to make do with what I had.

The Marconi installation is shown in **Figure 1**. It would have been nice to get the flat top higher in the air, but that was impossible. Because I couldn't use a bigger antenna, I had to look elsewhere to improve my signal. Knowing that the transmitter power had to flow through the ground connection, this seemed a logical place to make an improvement in my signal. The ground consisted of the copper water system in the house plus one ground rod. Discussions with DXers on the band quickly convinced me that my ground wasn't as good as I'd hoped it would be. A lot of RF was being wasted in ground resistance. To monitor improvements, if any, I placed an RF ammeter in series with the antenna. With a power output of 80 watts, I logged 1.27-A antenna current. Ohm's law showed my antenna feedpoint resistance was about 49.6 ohms — a good match to my transceiver, even if DX performance was unimpressive.

Improving the ground connection

I suspected that a lot of my output power wasn't going into the ground connection. Where else could it go? Perhaps it was going down the line cord and into the house wiring. Acting on this supposition, I wrapped the line cord around a ferrite rod and noticed the antenna current had now increased to 1.65 A. The feedpoint resistance of the antenna had dropped to 29.4 ohms. That indicated less ground loss. But now it was more difficult to match the antenna to the transceiver. I needed an antenna matching unit to achieve a 50-ohm interface.

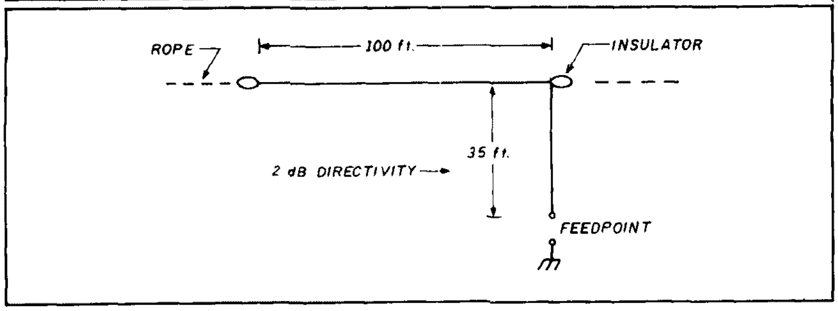
My next step was to add two quarter-wave radials to the system. These wires ran about a foot above the ground and wound in and out through the shrubbery in the yard. It was the best I could do. Unfortunately, there was no room to add additional radials. I had to be content with what I had. The radials brought the antenna current up to 1.82 A. This was another step in the right direction. Now the computed feedpoint resistance of the antenna was about 24.2 ohms.

There seemed to be a modest improvement in the antenna. I now found I could work some DX. Mine wasn't the loudest signal on the band, but the little antenna provided a lot of fun when the DX guns were occupied elsewhere. However, a nagging thought remained in the back of my mind. How efficient was the antenna? Had I really conquered the ground loss problem?

Computer analysis of the antenna

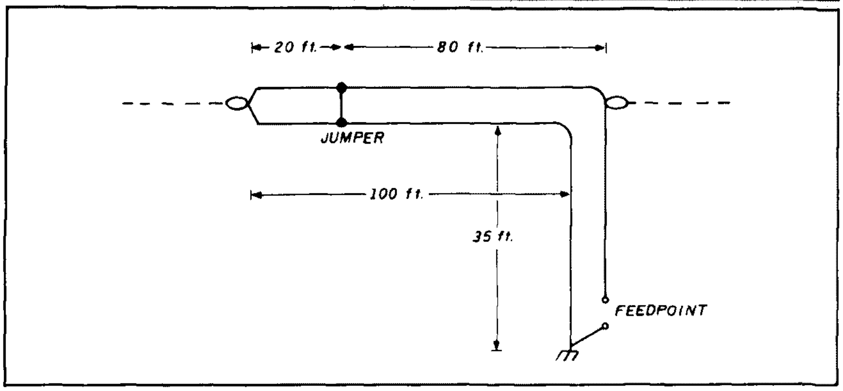
I didn't do much more with the antenna installation, and during the warm summer months I lost interest in 160-meter operation. But last fall I decided to get back on the band. Now I was able to analyze the antenna with the new K6STI computer program discussed in my last column.[1] The analysis revealed that the true feedpoint resistance of my antenna over typical soil in this location was only 7.8 ohms! Because my measurement of antenna current indicated a feedpoint resistance of about 24.2 ohms, the inescapable

FIGURE 1



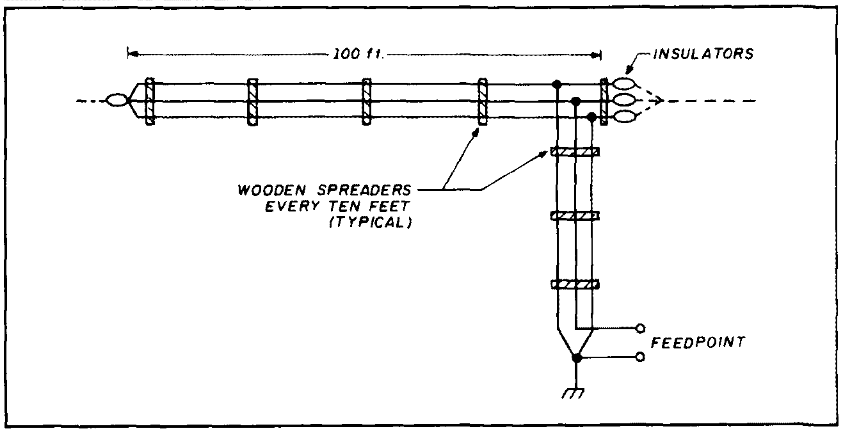
"Old Faithful" 160-meter Marconi antenna is full of surprises that will fool the unwary user. Antenna is self-resonant at 1.85 MHz.

FIGURE 2



"Twin-lead" Marconi antenna. TV-type 300-ohm line is shorted 20 feet from end to account for velocity factor of line.

FIGURE 3



Three-wire Marconi provides high value of feedpoint resistance. If outer wires had half the diameter of the inner fed wire, the impedance step up would have been 9:1.

conclusion was that the difference between the two figures (16.4 ohms) represented *ground loss*. My antenna efficiency was only about 32 percent! That meant that out of 80 watts, I was radiating only about 25 watts. The rest of my power (55 watts) was just warming the ground.

This also meant that I suffered about 4-dB signal loss in reception. No wonder I couldn't hear some of the weak DX on the band!

The solution: the folded Marconi antenna

In early spring I decided to change the single wire Marconi antenna to a folded system using two parallel wires (**Figure 2**). One wire was fed and the other returned to the ground connection. This provided an impedance step-up of four, and allowed an improved value of radiation resistance of 7.8 × 4, or 31.2 ohms. The ground resistance remained the same value as before (16.4 ohms), indicating that antenna efficiency was now about 66 percent. I had picked up 3 dB in transmitted power and had gained 3 dB in signal reception! Not a bad improvement for substituting 300-ohm ribbon line for the no. 14 wire in the antenna.

My final idea was to go to a three-wire folded antenna, which would provide an impedance step-up of about 11:1 (**Figure 3**). The antenna feedpoint resistance would then be 7.8 × 11, or 85.8 ohms. With 16.4 ohms ground resistance (which I didn't seem to be able to eliminate), the overall antenna efficiency rose to 84 percent! The small series of improvements gave me nearly 4-dB boost in transmission and reception at little cost.*

*That's an increase in power ratio of 2.51:1. Ed.

As I had no three-wire conductor on hand, I made one out of three no. 14 wires spaced 1 inch apart with a number of 4 inch long wooden spreaders. The assembly was a rat's nest on the ground but it straightened out when I got it up in the air and under tension.

A final run of the computer program showed that the Marconi exhibited about 3-dB directivity in the direction of feed, as Marconi had predicted long ago. In addition, the computer showed that the greater the length of the vertical portion of the Marconi, as compared with the horizontal section, the *greater the feedpoint resistance*. The limit, of course, is when the whole antenna is vertical. A single conductor has a feedpoint resistance of about 37 ohms (**Figure 4**).

How were my operating results? Much, much improved over the original design. At times, I even had DX stations answer my CQ. WOW!

Running the MN antenna program

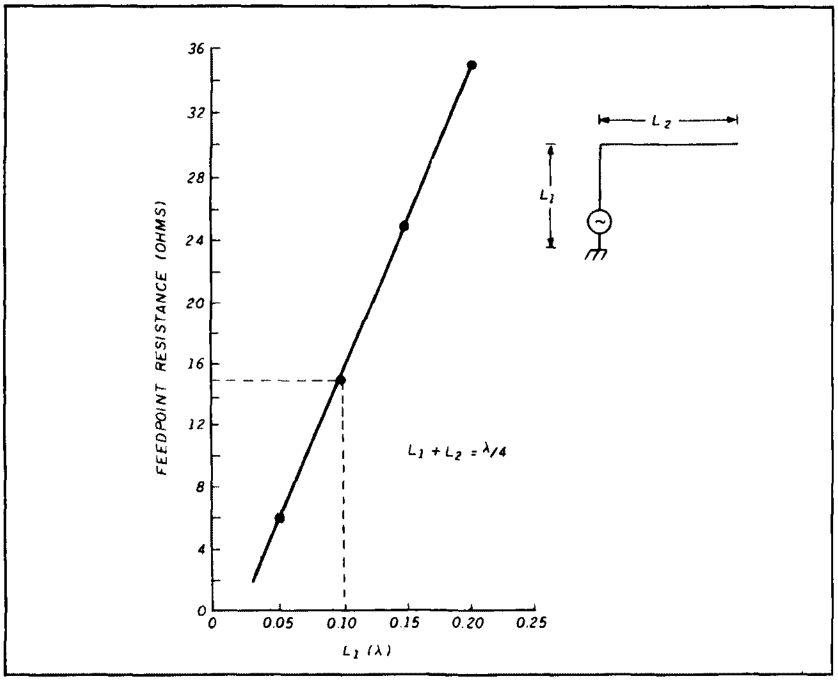
Last month I explained the technique of preparing antenna data for inclusion in K6STI's antenna analysis program.[1] The discussion covered program theory, the coordinate system used, and wire, segment pulses, and sources. All of this information is placed in a special format and input to the computer. The antenna in question may be modeled in free space or above ground. The program lets you specify ground conductivity in the area of the antenna. I model antennas in free space as the program runs more quickly. Elevation plots for antennas above ground are available in many handbooks. However, the books can't show the effect of lossy ground on the elevation patterns. This can be very pronounced, and isn't often something you can figure out intuitively. Thus, after the antenna is modeled in free space with satisfactory results, it's a good idea to run it again over simulated ground. Your local conductivity factor will give you an insight into the reflection gain.

Modeling a sample antenna

As a working example, I'll discuss a popular 20-meter, three-element beam. The design is shown in **Figure 5**. The antenna is built on a 22-foot boom. Elements are assumed to be 1 inch in diameter. (Element diameter and taper will be discussed in my next column.)

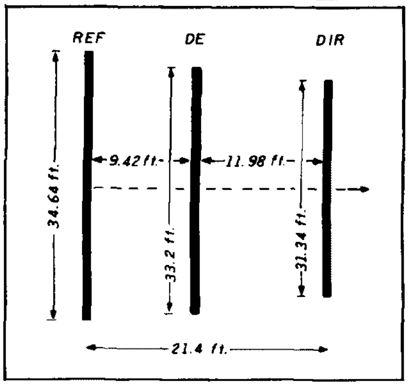
While linear element dimensions are

FIGURE 4



Bent-wire Marconi has low feedpoint resistance, depending on ratio of L_1 to L_2. When L_1 is 0.1 wavelength, for example, feedpoint resistance is about 15 ohms.

FIGURE 5



Three-element 20-meter Yagi. Design frequency is 14.175 MHz.

shown, the MN program requires information in a different form. The tip position of all elements is expressed in X-Y-Z Cartesian coordinates. (Because the antenna is only two dimensional, the Z coordinates are zero.) The appropriate X-Y-Z data for this beam, plus other required information, are shown in **Table** 1.

To create an antenna file, you need a text editor or word processor. The MN package includes a text editor called TED. This is a short program and works much in the manner of WordStar™. You can use other programs, like EDLIN, if you wish. But TED does the job quickly and easily.

Once the information has been placed in the word processor in the proper form, give the data a name and an extension (.ANT), and the file is entered into the antenna program. In this case, the name used is 3LYAGI.ANT.

After you view the antenna file for accuracy, start the computation to determine the gain, front-to-back ratio, and input resistance (impedance) by giving the command "G". In this example (using a math coprocessor) the matrix fill time is 21 seconds. The computed antenna information is shown in **Table 2**.

This particular antenna is designed for good gain with a high degree of front-to-back ratio. About 1 dB of the maximum possible gain is sacrificed to achieve this favorable ratio. Input impedance is good, with the driven element being slightly short for the design frequency of 14.175 MHz.

Note that my printer used an italic "x" in place of the percentage sign in the beamwidth printout. IBM clones and printers occasionally have little idiosyncracies like this and some diddling with the dip switches is required (sigh).

The last step in the program is to display the directive pattern. If you want to print it, you'll need a dot matrix printer (see **Figure 6**). And there you have it! All of this vital data is derived without cutting a single piece of aluminum or climbing even one foot up a tower!

The antenna file

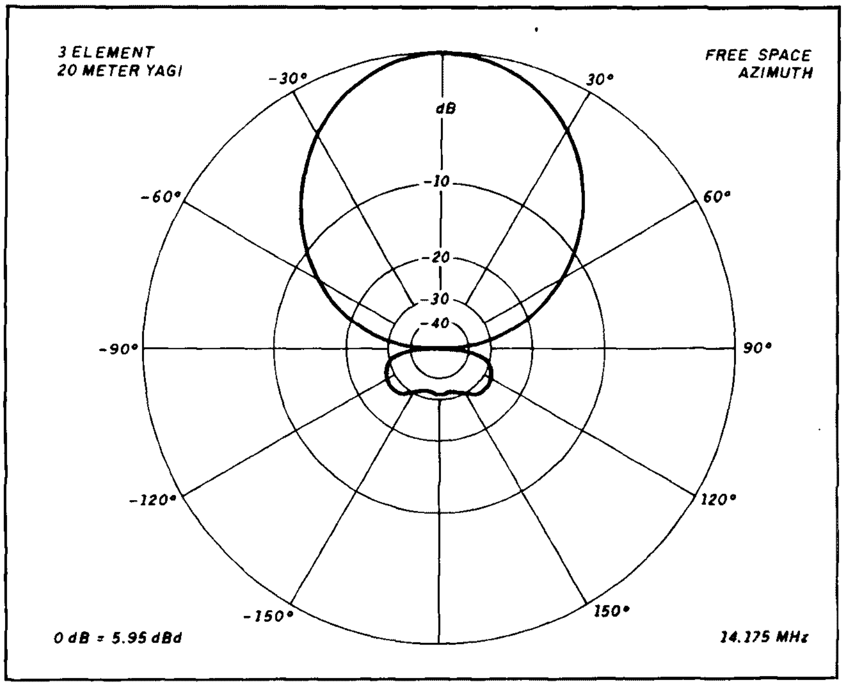
The MN program has a library of many different antenna files which you can examine before you input your own design. It's a good idea to examine these files to get the "feel" of how the program works and how the data is input to the program. Start with a simple antenna, like a dipole or two-element Yagi. Note how the X-Y-Z coordinates are derived and pay attention to the spaces between statements, letters, and numbers within the individual lines. If results seem odd, examine the RUN file to see if you have allocated pulses and the source properly. You'll find that reviewing the antennas in the file, before you run your own program, pays big dividends. MN has built-in prompts which lead you step by step to the conclusion. You may be dismayed at some of the ego-deflating gain figures for beam antennas

TABLE 1

Program for 20-meter beam. See *Reference 1* for explanation of entries.

```
Three-element 20-Meter Yagi
14.175 MHz
Free Space
3 wires, feet
10   -10.7,-17.32,0  -10.7,17.32,0        0.083
10   -1.28,-16.6,0   -1.28,16.6,0         0.083
10    10.7,-15.67,0   10.7,15.67,0        0.083
1 source
14
0
```

FIGURE 6



Azimuth pattern of 20-meter Yagi combines good gain with excellent front-to-back ratio.

TABLE 2

Twenty-meter beam has 5.95 dBd gain, over 31 dB front-to-back ratio and feed-point impedance of about 22 ohms. (See text regarding beamwidth printout.)

Matrix Fill Time	0:21
Matrix Factor Time	0:02
Impedance	22.3 −j 3.9 ohms
SWR	2.26 for Z = 50 ohms
Forward Gain	5.95 dBd
F/B	31.83 dB
Maximum Sidelobe	26.98 dB down at 124x Azimuth
Azimuth Beamwidth	63x
Elevation Beamwidth	100x

provided by MN which contradict highly touted figures given by some beam antenna manufacturers, but that's the way it is.

Next month I'll review the Yagi optimizer program which lets you manipulate the dimensions of your antenna and see what happens when you change length, spacing, and taper of Yagi antennas. Stay tuned!

The Dead Band Quiz

Thanks to the following additional readers who responded to my problem about the snowplow. (See last month's column for the solution.) They are: AC5P, KE2MO, N6SVI, NG5F, WY7U/4, James Conley (no call given), W6MUR, WA7HVT, and K4KQS.

Some readers have requested a "literary quiz" instead of a mathematical-type problem. So here are two little quizzes to whet your appetite:

"Remain on patrol in vicinity of Rockall."

This unusual signal was sent by whom to whom? What was the approximate date the signal was sent, and what was the significance? What, or who, is Rockall? What is the story and who is the author?

The second quiz concerns a popular TV show now in rerun. Who said the following and under what circumstances?

"Brain! Brain! What is brain?"

Good luck and see you down the log. Written replies to these little brain teasers will be acknowledged in this column. My QTH: Box 7805, Menlo Park, California 94025.

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DIGITAL VOICE STORAGE
IN THE HAM SHACK

Digital voice storage on the IBM PC

By Carl Lyster, WA4ADG, 4412 Damas Road, Knoxville, Tennessee 37921

In my February 1989 *Ham Radio* article,[1] I described the basic principles of a digital voice storage playback only device suitable for use as a repeater IDer. This month I'll discuss the hardware and software required to add digital voice storage capability to any of the IBM series of PCs or clones. In addition to giving you the ability to store and play back audio in the range of 1 Hz to 5 kHz, this project lets you program audio passages in ROM for use with the voice IDer.

This device is the result of years of tinkering with A/D and D/A converters attached to microcomputers. My initial work was done wih a South West Technical Products 6800 computer. I then progressed through PDP-11/34s to the IBM PC. I made each transition in search of more memory in an effort to lengthen the storage time. The IBM, with its 640K of memory and bussed card slots, is a natural choice for hardware experimentation.

In 1987 I packed an IBM PC along with a gas generator and an earlier version of this card to the Dayton Hamvention™ *flea market*. I was overwhelmed by the interest shown and suggestions of possible uses for this concept. It has taken a year of reworking the design to make it easily reproducible and I hope it may form a de facto standard for Amateur Radio digital voice on the IBM. The applications for this technology are limitless and can be put to immediate use by hams in such areas as aids to the blind, simplex repeaters, and voice mailboxes. I'm hoping this article will inspire professional programmers to create some sophisticated software packages.

Description

This card, when installed in an IBM or clone, allows the digitization of human speech, storage of the passage on hard or floppy disk, editing of the digitized voice, and playback through a suitable audio amplifier or transmitter. A frequency response of 5 kHz gives a fidelity comparable to a narrowband FM 2-meter signal. The digitization rate is 10K samples per second set by software timing loops and therefore changeable by keyboard command. At this standard rate of 10K samples/second a 640K IBM CPU will store about 46 seconds of continuous voice. Some memory is used by DOS, the BASIC interpreter, and the digitizing program itself. I chose to use the BASIC language because it handles binary files easily and is almost universally popular.

Technical information

My hardware takes advantage of some clever design work by the people at IBM. They realized the need to dedicate a small portion of the I/O address space available in the PC to "hardware hacker" types like me. This I/O area lies in the range of 300 hex to 31F hex and is given to the whim of the computer user. Any user-supplied device operating in this I/O area should be free of interference to or from the computer. Naturally, several mail order electronics firms realized this potential and have developed easy-to-use prototyping cards which operate in this user I/O area. To date, my work has been done with a prototyping card called the PR2, sold by JDR Microdevices. I recommend their card for this project and any other which requires bus interfacing to the IBM.

Hardware description

Your project can take one of two forms. First, you may purchase a PR2 card and hand assemble the circuit (see **Figure 1**) on the card using its on-board logic. This reduces the complexity somewhat but results in greater expense. Second, you may follow the pc board artwork I have provided (see **Figures 2** and **3**). It contains all of the elec-

tronics on a one-third length double-sided pc board. This is the simplest and most compact method and will provide better rejection of computer-generated noise. I've built many units using both types of construction and have been pleased with all of them.

The digitization and playback is performed by a pair of ICs produced by analog devices. I used the AD 7224 D/A converter for playback and the AD 7574 A/D converter for recording. These are eight-bit converters which fit the addressing scheme of the IBM nicely (one data point or sample equals eight bits or one byte). The dynamic range of the recording is approximately 48 dB with a frequency response of 1 Hz to 5 kHz.

Signal flow for the one-third length card is as follows. During recording, audio enters the card at a maximum level of 0.5 volts p-p and is amplified and level shifted by op amp IC1. The gain is set to 10 by R1 and R2, and the DC offset or "bias" of the amplified signal is adjusted by pot VR1 to 2.5 volts DC with no audio present. The resulting level-shifted audio will swing from 0 to 5 volts for an input of 0.5 volts p-p. Under software timing, a machine language write to I/O port 304 hex starts the A/D conversion sequence.

The A/D IC-11 converter responds by lowering its busy flag (pin 14), which causes IC2 (a sample and hold chip) to store temporarily the present amplitude of the audio signal. A sample and hold chip is analogous to a capacitor that can be turned on or off by a TTL level (the busy flag). The stored output of the sample and hold chip is then filtered by an eighth order equal component Sallen-Key low pass filter. I described the operation of this filter in my February 1989 article, so I won't go into detail here. After about 20 μs the A/D converter finishes digitizing the sampled audio and clears its busy flag. The software timing loop then issues a read command to I/O port 304 hex. This retrieves the digitized data from the A/D converter and stores it in the CPU, which transfers the data byte to RAM memory. The timing loop restarts the entire process 1/10,000 of a second later for the next data point. When the desired RAM space has been filled, the software program halts the recording process.

The digitized passage is reproduced when the CPU retrieves a data byte from RAM and sends it via a machine language write to I/O port 308 hex, D/A converter IC10. The AD 7224 digital to analog converter produces an output voltage proportional to the magnitude of the binary number it is sent. Binary 00000000=0 volt DC, binary 10000000=2.5 volts DC, and binary 11111111=5 volts DC. The analog voltage output is then filtered by an eighth order Sallen-Key low pass filter, identical to the one in the recording unit, to remove any frequency components above 5 kHz that would cause aliasing. The reproduced audio is sent off the card to an external audio amplifier or transmitter. Software timing loops instruct the CPU to retrieve the next data point from RAM 1/10,000 of a second later and repeat the process until the desired time has elapsed.

IC 7, 8, and 9 form an I/O port decoder which operates in the user I/O area. Pin 14 of the decoder (IC9) goes low for address 304 hex to 307 hex to select the A/D converter; pin 13 goes low for address 308 hex to 30B hex to select the D/A converter. I didn't provide external bus buffering because both converters contain on-chip buffering, which has proven to be sufficient for the most heavily populated motherboards. You can obtain power for this card from the computer's busses, but my experience has shown them to be too noisy for low level audio use. Because of this, I use two voltage regulators and two zener diodes to ensure ripple free power. Raw $\pm$ 12 volt supplies from the computer's card slot are first regulated down to $\pm$ 8 volts for the low pass filters and the sample and hold chip. The 8-volt supplies are then regulated down by zener diodes to $\pm$ 5 volt supplies that are used as references for the converter chips. The CPU's bulk 5-volt logic supply powers the Vcc requirements of the voice card's ICs directly.

Those who elect to build their project on a PR2 card won't need to assemble the I/O port decoder because the PR2 contains nearly identical electronics.

Checkout and adjustment

After you've built the card and inserted it into the computer, you can run two short programs* to verify proper operation and adjustment of pot VR1. Program ADTEST.BAS (Listing 1) is used to adjust VR1. This program is loaded under BASIC just like any other. It samples the voltage present at the audio input jack continuously and displays the magnitude, in decimal, of the converted data point. With no audio applied to the card, adjust pot VR1 to display a value of 128 $\pm$ 1 on the screen. This corresponds to 2.5 volts DC. After you've adjusted VR1 correctly, connect an audio source of 0.5 volts p-p to the card and run ADTEST.BAS again. Data point values ranging between 0 and 255 should appear randomly on the screen. This program samples the A/D converter much too slowly to store audio; it is for test purposes only.

DATEST.BAS (Listing 2) checks the D/A converter for proper operation. This program generates a slow saw-toothed waveform that can be observed with a scope or meter movement type multimeter. A clean ramp from 0 to 5 volts DC on the scope or a smoothly moving needle on the multimeter indicates a functional D/A converter and low pass filter.

Please keep in mind that the audio output of this card is DC coupled and may need to be isolated from your external device by a blocking capacitor of about 5 μF.

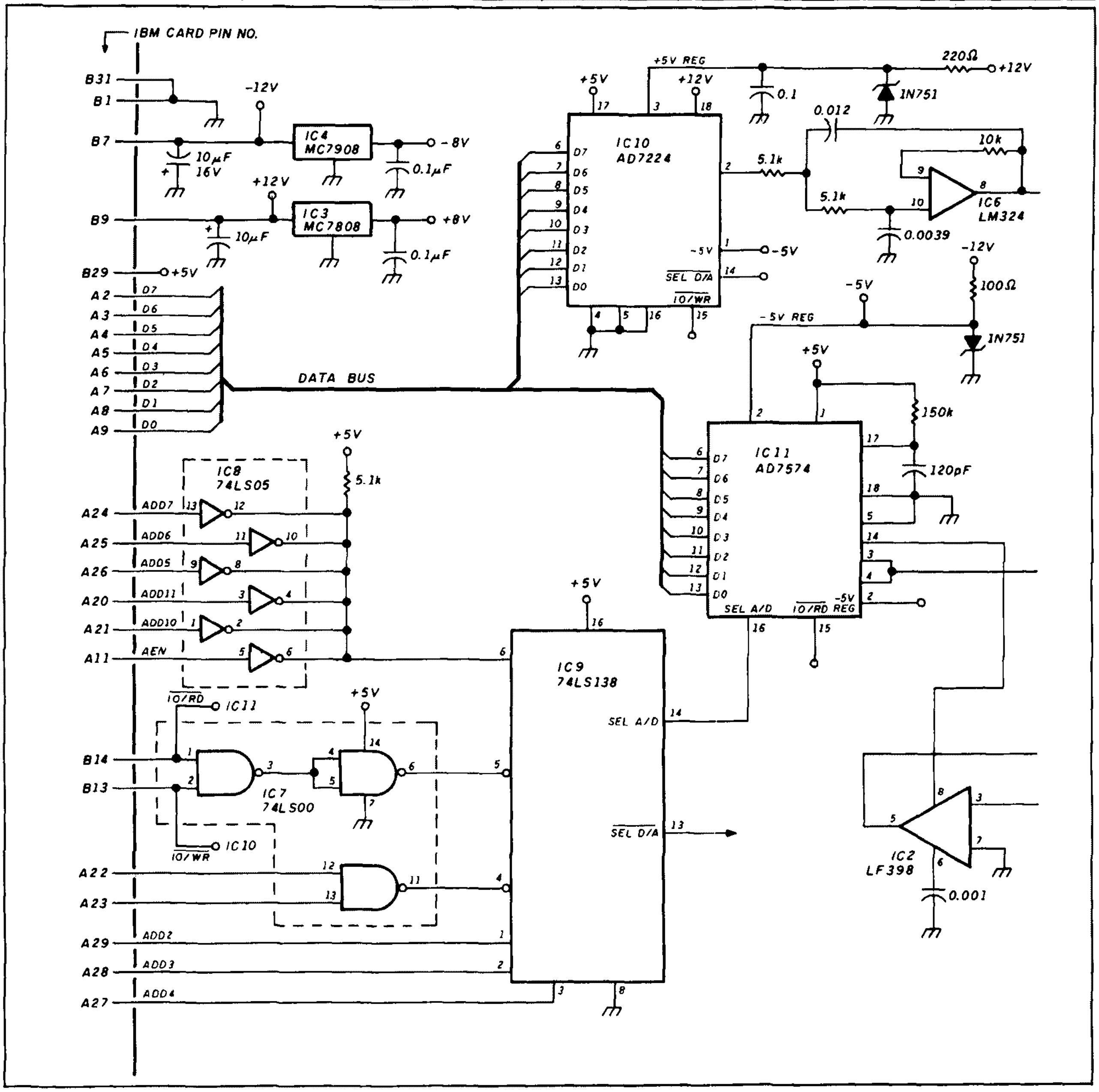
Operating software

The software to perform the voice digitization and playback consists of two machine language subroutines (Listings 3 and 4) that are "poked" by BASIC into RAM memory. The bulk of the BASIC program, HAMTALK (Listing 5), is used to save and retrieve binary files containing the digitized audio. My program is limited due to the lack of space and a desire to spare you from having to enter several hundred lines of code. It does provide the ability to save and retrieve voice on disk, alter the sampling rate for experimentation purposes, and program ROM for the voice IDer. Because of the simplicity of this program, it will only store up to 25 seconds of voice. A more sophisticated program that makes use of the full memory capacity of the IBM and allows editing of the recording in memory is available from me on two 5.25-inch floppy disks.

The program HAMTALK.BAS will operate with GWBASIC or BASICA. When starting BASIC, you must use the /M switch to set aside RAM space for the machine language subroutines. Do so by typing BASICA /M:15000 and, after receiving the BASIC "OK" prompt, starting the program by

* Listings 1-5 are available from *Ham Radio* for an SASE with $.25 postage.

FIGURE 1



Circuit diagram.

typing RUN HAMTALK. The screen will clear and a menu will appear. I've set aside 256K of memory for the voice buffer. This corresponds to about 25 seconds of continuous voice. *Your machine must have at least 512K of RAM to operate. Any less will result in your voice buffer overwriting DOS, which will hopelessly lock up your computer.* A short summary of the menu choices follows.

R — The R command starts the recording of 256K worth of memory, normally about 25 seconds.

P — The P command starts the playback of the 256K voice buffer.

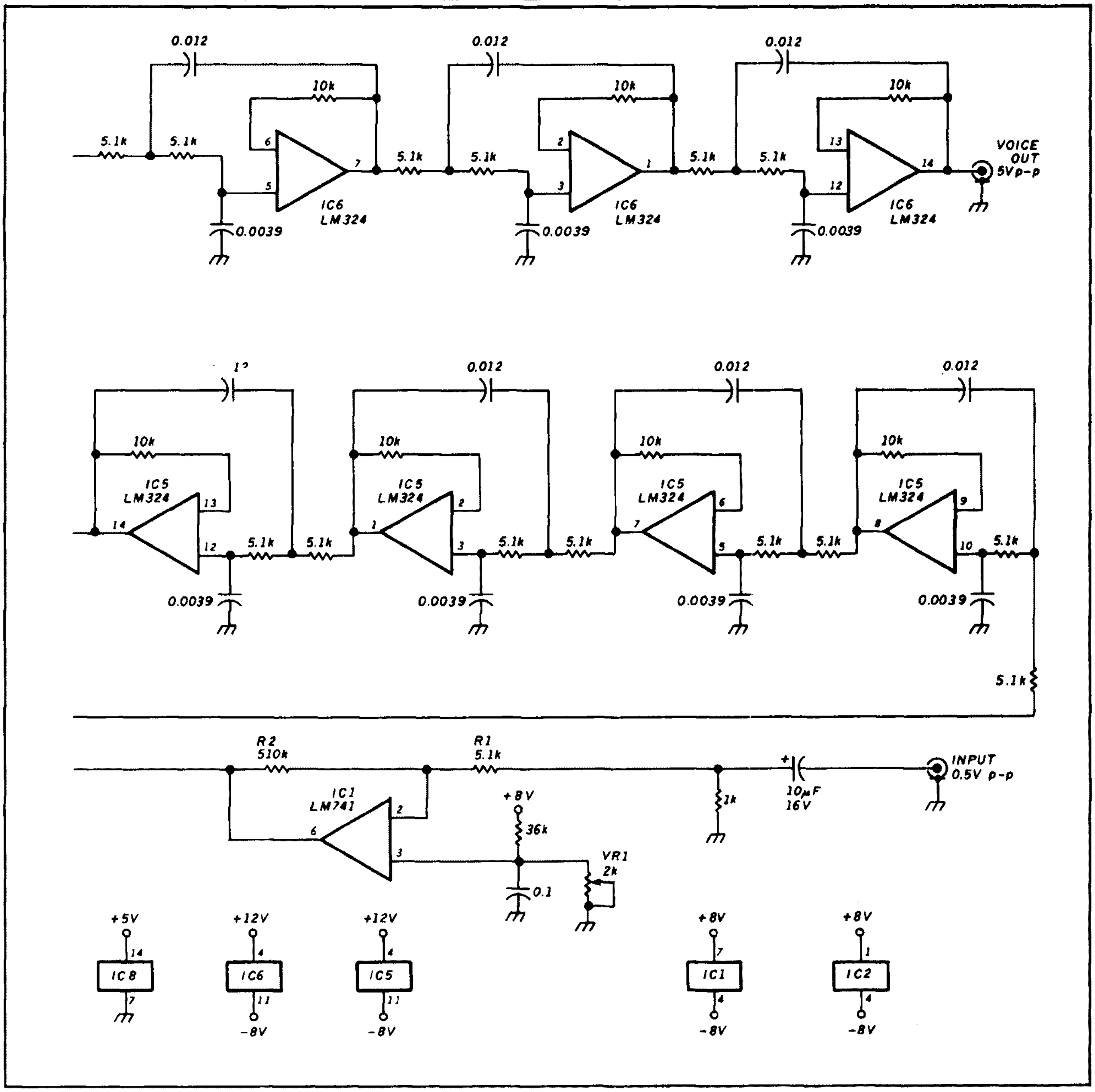
D — The D command allows keyboard selection of the digital sampling rate. The default value is 11, which corresponds to 10K samples per second on a 4.77-MHz machine. If your PC operates faster than 4.77 MHz, you'll have to increase D proportionately.

Q — The Q command returns you to DOS.

DIR — The DIR command displays a directory of the voice files contained on the default disk drive (the drive where BASICA resides).

S — The S command saves the voice buffer to disk. The buffer is composed of four files of 64K. Each file is 6.4 seconds long. If you wish to save the first 12

FIGURE 1



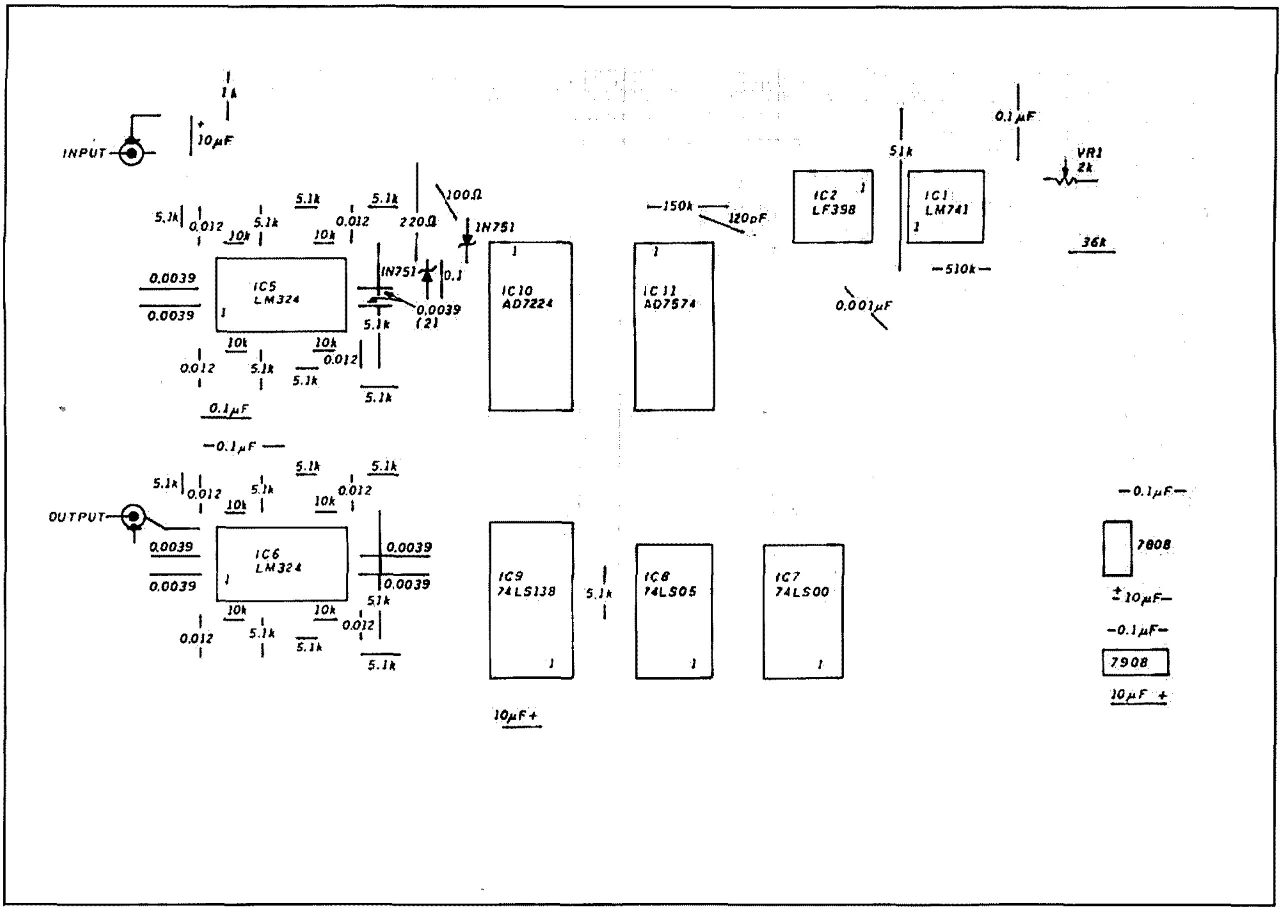
seconds of the buffer, you must save two files. Three files would save the first 19 seconds and four files would save the entire 25-second buffer. You'll be prompted to enter the number of files to save and then asked for a name to store the files under. For instance, if you enter four files with the name TEST, the program will save your buffer as four files with the names TEST1.BAS, TEST2.BAS, TEST3.BAS, and TEST4.BAS. Each file will represent one 64K segment of the voice buffer.

G The G command retrieves voice files from disk and stores them in the voice buffer. You are prompted for the number of files you want to retrieve and the file names.

Hard drive notes

I strongly recommend that you use a hard drive. There's a quantum leap in file access time between a floppy and a hard disk. If you place your software on a hard disk, *do not* store the voice program or voice files in a subdirectory; place them in the root directory. I have found that the use of subdirectories slows the access time of the hard disk by a factor of 4.

FIGURE 2



PC board artwork, component side.

PARTS LIST

Capacitors

1	120-pF	50-volts DC	disc
8	0.0039-μF	50-volts DC	metalized film, Digi-Key P4558
1	0.001-μF	50-volts DC	disc
8	0.012-μF	50-volts DC	metallized film, Digi-Key P4514
4	0.1-μF	50-volts DC	disc
3	10-μF	16-volt	electrolytic (radial)

Resistors

1	100 ohm, 1/4 watt
1	220 ohm, 1/4 watt
1	1 k, 1/4 watt
17	5.1 k, 1/4 watt
9	10 k, 1/4 watt
1	36 k, 1/4 watt
1	51 k, 1/4 watt
1	150 k, 1/4 watt
1	510 k, 1/4 watt

Potentiometer

1	2-k trimpot, Digi-Key D1AA23

Semiconductors

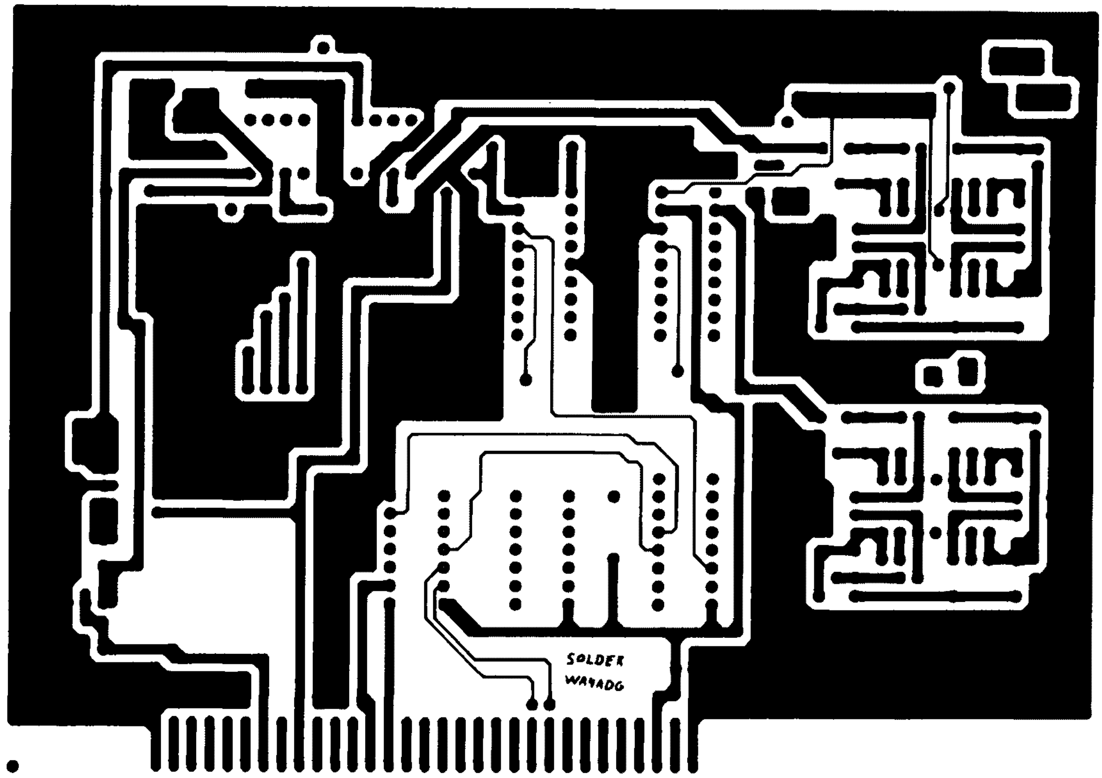
2	1N751 5-volt zener
2	LM324 quad op amp
1	LF398 sample and hold amp
1	LM741 high performance op amp
1	MC7808 three term, +8 volt, 1-A regulator
1	MC7908 three term, −8 volt, 1-A regulator
1	74LS00 quad 2 input NAND gate
1	74LS05 hex inverter
1	74LS138 expandable 3/8 decoder
1	AD7224* D/A converter
1	AD7574* A/D converter
1	printed circuit board**

* Available from Allied Electronics and Pioneer Standard Electronics outlets nationwide.
** Available from FAR Circuits, 18N640 Field Court, Dundee, Illinois 60118 for $15 each. Boards *do not* have plated-through holes and component leads must be soldered on each side of the pc board.

Mathematical theorems — hearing is believing!

In my first article, I gave a brief description of voice storage and the theories of digital sampling. You may recall the Nyquist theorem stated that a sine wave must be sampled at least twice per cycle in order to be reproducible. This project can convey the reality of the Nyquist theorem far better than any text book. I stated earlier that the sampling rate will be 10 kHz with a 4.77-MHz CPU and a "D" of 11. With a low pass filter response of 5 KHz, a D of 11 is absolutely the slowest sampling rate that falls within the rules of the Nyquist theorem. When the D value is changed to 12, which slows the sampling rate to about 9.5 kHz, you can literally hear the reproduced audio "falling apart." This is characterized by a ringing sensation that's the result of misreproduced frequencies. I have provided the ability to change the sampling rate just for this purpose, to illustrate graphically a rule which we are becoming more dependent upon each day.

FIGURE 3



PC board artwork, trace side.

Programming EPROMS

HAMTALK.BAS stores voice files in binary format. These files can be read by an EPROM programmer, sold by JDR Microdevices. The programmer that I've been using was sold under the name "Sunshine Programmer." It has been replaced by a new version that costs about $130. If you own a different EPROM burner, you may have to be creative in order to get the data from disk into your particular programmer. You might try moving memory blocks with machine language subroutines to an area of memory used by your programmer or dumping the disk file out of a serial port. In any case, once the data is in your programmer, simply program the appropriate chip (being careful to use the correct voltage), then place the EPROM in your voice storage unit.

If there's sufficient interest, I'll design a low cost programmer to fit the IBM bus and read the voice files and program 27256 or 27512 chips in BASIC. Please write and let me know if this is something that you're interested in.

Program subroutines for voice digitization and playback

You can add voice capability to your favorite program once you understand the working principles of the record and playback subroutines. Both routines are written in machine relocatable code. This means they can be placed anywhere in RAM memory. You must take care to ensure that the RAM where these routines are placed is truly free and won't be used by other programs. Both playback and record routines contain starting and stopping addresses for the voice buffer, as well as a delay number to control the sampling rate.

To use the record routine, first set the starting and stopping addresses in hex of the range of RAM you wish to fill. In BASIC language, you'd POKE these values into the subroutines. If you wish to change the sampling rate, you'll also have to POKE a new value for the delay factor. This wouldn't normally be necessary. Keep in mind that a designed sample rate of 10 kHz will fill FFFF hex locations in 6.4 seconds. After setting the address, simply use a call subroutine statement to start the digitization. When the subroutine finishes digitizing the desired amount of RAM, a "return far" (RETF) instruction returns control to your program. The voice buffer is now filled and can be saved to disk or reproduced by the playback routine. The playback subroutine is used in much the same way as the record routine. Set the starting and ending addresses of the voice buffer and then call the

subroutine. It is assumed that your program has either previously digitized a passage into the voice buffer or filled the buffer from disk. The RETF command returns control to your program after the voice passage has been reproduced.

Before I present a detailed operation of the subroutines, a short simplified description of how the IBM addresses its RAM will eliminate a great deal of confusion. A common CPU, like the old 8080, addresses memory with 16 binary address lines A0 to A15. This gives 65,536 separate addressable bytes, 0 to FFFF hex. The 8088 can address 1,048,576 bytes directly using address lines A0 to A19, corresponding to 0 to FFFFF hex. The addressing confusion arises when you try to store a 20-bit wide address in a 16-bit wide register! Intel has solved the problem using a segment register and an offset register combined to produce a 20-bit wide address. For the purposes of this software and to keep life simple, consider the segment register to be nothing more than a bank select register which allows you to choose one of 16 64K byte banks. The offset register contains the address location in the currently selected 64K bank. The correct method of describing a memory address using this scheme would look like this: segment address to the left of the colon, offset address to the right of the colon. One example might be 1000:3FFF, which corresponds to bank 1 offset 3FFF. For your purposes, the segment register or "bank select" register will only contain one of 16 values. These are in hex and range from 0000 to F000. The first digit is the "bank number" and the 000 portion is necessary filler. The banks are tabulated below.

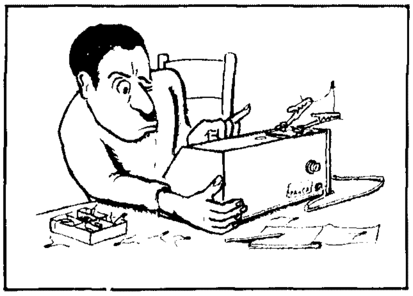
0000:= 1st 64K bank	8000:= 9th 64K bank
1000:= 2nd 64K bank	9000:=10th 64K bank
2000:= 3rd 64K bank	A000:=11th 64K bank
3000:= 4th 64K bank	B000:=12th 64K bank
4000:= 5th 64K bank	C000:=13th 64K bank
5000:= 6th 64K bank	D000:=14th 64K bank
6000:= 7th 64K bank	E000:=15th 64K bank
7000:= 8th 64K bank	F000:=16th 64K bank

This scheme forms a simple method for the voice routines to access the entire RAM space of the CPU. When one 64K byte bank is full, simply increment the segment register by 1000 hex to select the next 64K byte bank. In reality, the segment:offset operation of the CPU is much more complicated than this brief description.

One last important point. Because the timing loops for the digitization are software generated, all processing by the CPU must be halted during the voice routine operation; this includes the keyboard input! To accomplish this, each voice routine turns off all possible interrupts while running and then enables interrupts at completion of the routine. Both routines load the voice buffer starting address with an MOV BX,0000 for the offset address and an MOV DI,2000 for the bank address. This equates to byte zero of the third bank — just above and out of the way of BASIC. The ending addresses are loaded as part of compare instructions like CMP BX,0000 for the offset register and CMP DI,6000 for the segment register. You may change these starting and stopping points to any location in RAM as long as you don't overwrite a resident program, like DOS!

Next time I'll discuss combining a clone motherboard with this month's project, along with ROM or RAM based software, into a digital voice storage simplex repeater suitable for whatever RF frequency you choose. hr

Joseph J. Carr, K4IPV



ANALOG METER MOVEMENTS: HOW TO USE THEM

Although the digital meter has popped up all over the electronics industry, it's noteworthy that a lot of Amateur Radio equipment, both factory made and homebrew, still uses the analog meter movement. At the last large hamfest I attended, I saw that analog meters were popular among those who pored over the tailgaters' offerings.

Why, in an age of digital everything, is the analog meter still popular? I think there are two basic reasons. First, the analog meter isn't terribly sensitive to RF fields that surround Amateur Radio gear. Second, this meter is most often used in Amateur Radio to find peaks and dips rather than an actual value. While the digital meter has a certain edge over analog types when it comes to reading values with ease, its very nature makes looking for peaks and dips annoying — especially if the integration of the digital meter is long!

In this month's column I'll look at analog meters and their applications. While I don't expect to exhaust the field, this information should be useful for a wide variety of Amateur Radio metering applications. I'll examine the basic DC meter movements and a method or two for making them read AC values.

DC instruments

The two most common forms of DC meter movement are the *D'Arsonval* and *taut band* designs. Both movements are examples of a general class called *permanent magnet moving coil* (PMMC) *galvanometers*. These devices work on the same basic principle as the DC motor. A simplified view of the PMMC movement is shown in **Figures 1A** and **1B**. A movable coil is mounted in the magnetic field between the two poles of a permanent magnet. A current flowing in the wire generates a magnetic field. The polarity of the magnetic field is determined by the direction of the current flow, while the strength of the field is determined by the current magnitude.

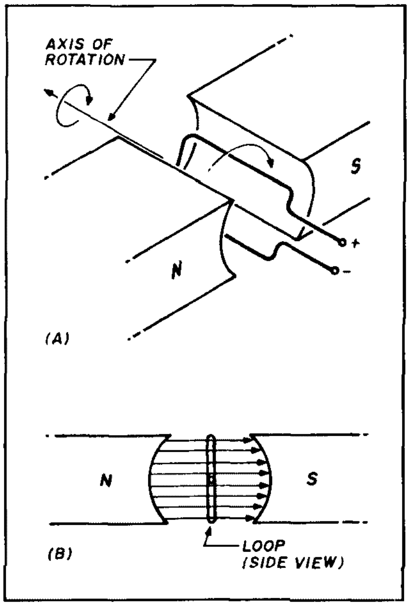
The coil of a PMMC movement is mounted so it can rotate in the space between the magnet poles. A current in the coil creates a magnetic field that either aids or opposes the field of the nearby magnet poles. Current flow in one direction causes a clockwise rotation; current flow in the opposite direction causes a counterclockwise rotation. The amount of rotational position change is proportional to the current magnitude.

The D'Arsonval meter movement

Figure 2 shows the D'Arsonval meter movement. A side view of the meter movement, without the permanent magnet, appears in **Figure 2A**; a front view of the magnet is shown in **Figure 2B**. The coil in **Figure 2A** is wound on an armature (or bobbin) which is mounted on a pair of jeweled bearings to reduce friction (see **Figure 2B**).

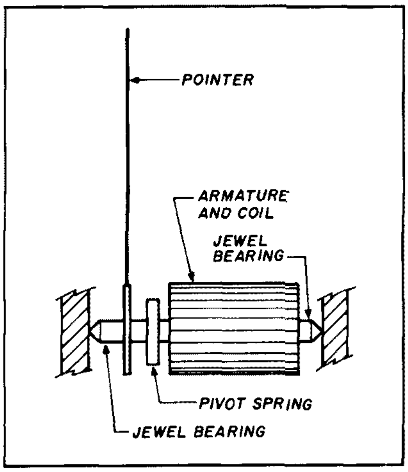
When a current flows in the coil, the armature assembly deflects clockwise (as illustrated in **Figure 2B**) by an amount proportional to the current strength. The amount of deflection can be marked in units of current on the

FIGURE 1



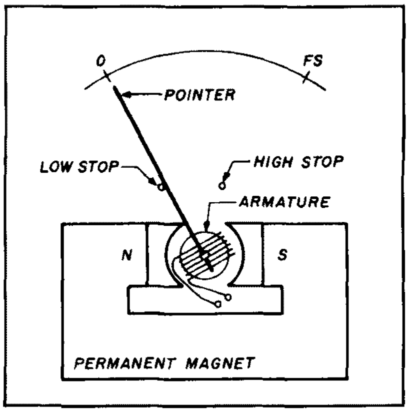
(A) Permanent magnet moving coil (PMMC) galvanometer. (B) Side view.

FIGURE 2A



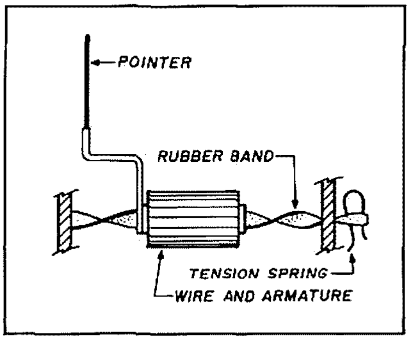
Side view of the D'Arsonval meter movement.

FIGURE 2B



End view of the movement.

FIGURE 3



Side view of the taut band meter movement.

dial scale. The coiled pivot spring dampens the pointer movement and returns the pointer to the zero position when the current flow in the coil ceases.

The travel of the pointer is limited by high and low end mechanical stops just beyond the zero and full scale limits printed on the dial scale. The wires to the coil are given just enough slack so that they won't be stretched anywhere in the pointer's normal range of travel. Overranging, however, can cause damage to both the pointer and the spring.

The taut band meter movement

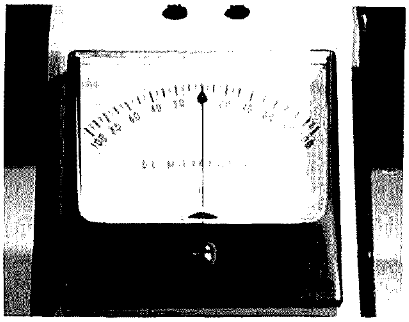
The taut band meter movement is essentially the same as the D'Arsonval movement, except for the manner in which the armature is mounted (see Figure 3). In the taut band meter movement, the armature is suspended from fixed supports on a stretched (taut) rubber band. The band is twisted as the armature rotates, so no restoring force from a pivot spring is needed.

There are two principal advantages of the taut band meter movement over the older D'Arsonval meter movement — greater sensitivity and durability.

Older D'Arsonval meter movements are rarely found with full scale deflection sensitivities less than 50 μA, but there are taut band models available that boast a full scale value of only 2 μA.

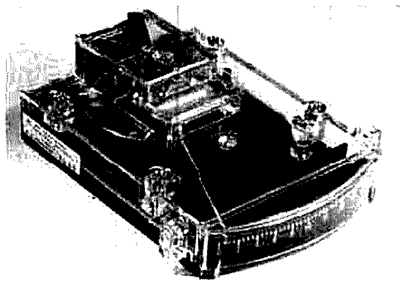
D'Arsonval meter movements are more easily damaged than taut band types because their jeweled bearings are more fragile. Even a short fall to the floor or bench top is sufficient in many cases to destroy the D'Arsonval meter movement (although some are notably more robust). The rubber band can snap in a taut band meter, but this occurs less frequently than does bearing damage.

PHOTO A



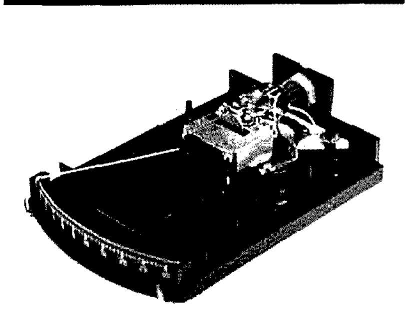
Zero center meter movement.

PHOTO B



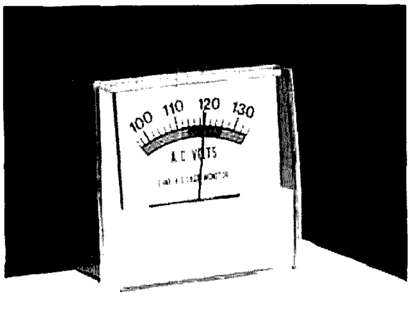
Edge mounted meter.

PHOTO C



Internal view of edge mounted meter.

PHOTO D



Expanded scale meter.

Types of analog meter movement

The two basic PMMC meter movements are available in a large array of sizes and types, but only a few can represent a wide spectrum of models. Meters can be classified according to the type of scale. The standard form shows zero on the left and full scale (FS) on the far right.

The terminals on the back of the meter indicate which terminal is positive. When the meter is connected to a circuit with the positive terminal to the positive side of the circuit, the deflection will be upscale when current flows. Reversing the connections forces the meter pointer backwards, and this may damage the meter.

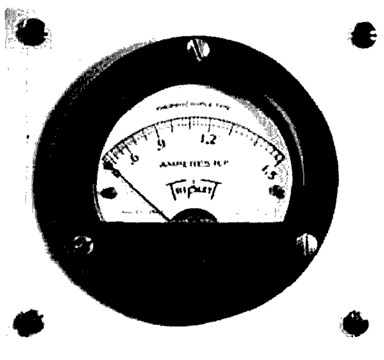
Photo A shows the zero center PMMC galvanometer movement. This scale has the zero point in the center. Positive values are to the right and negative values are to the left.

Photo B shows an edgewise meter movement. This design lets you conserve valuable panel space al the expense of increased depth. An interior view of this meter is shown in Photo C. Edgewise meters are available in both left and center zero models. The movement shown in Photos B and C has a pair of front panel tabs that can be used to set high and low limits for an alarm circuit. A pair of internal photocells and lamps are blinded when the meter pointer exceeds the set point. External circuitry can then detect the alarm condition.

Photo D shows an example of an expanded scale meter movement. The left-hand scale isn't zero, but has a voltage applied that's a little less than 100 volts AC. The expanded scale meter improves the meter's resolution in a range of interest. In Photo D the meter is used to monitor the AC power line voltage, which is normally constrained from 95 to 125 volts AC.

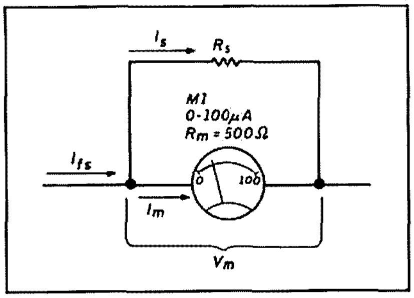
Photo E shows a meter that was once quite popular with Amateurs — the RF ammeter. This instrument has a left zero scale, but works differently from the normal PMMC instruments. Although a PMMC movement might be at the heart of this instrument, a thermocouple embedded in a resistive element inside the meter makes it work. Heat is generated when RF current flows in the resistive element. This heat causes a voltage to appear across the ends ol the thermocouple. The voltage is proportional to the RF

PHOTO E



Thermocouple RF ammeter.

FIGURE 4



Increasing current scale with a shunt resistor.

current, although you should be able to tell from the scale that it's a nonlinear function of RF current.

Using DC meters

There are three basic rules to remember when using DC current meters:

- Connect the meter in series with the load or circuit in which it is used.
- Use a meter that has a full scale current reading greater than the expected current flow.
- Use a meter that has a low internal resistance compared with the circuit in which it is being used.

A current meter is *always* connected in series with the load. Failure to observe this simple precaution can, and most likely will, result in total loss ol the meter. In most cases, the series requirement means that the circuit will have to be interrupted physically for the meter to be installed.

Apparent "errors" in meter readings

Assuming that your meter is in good shape, there's a possibility ol an error in your readings if the meter resistance, relative to the circuit resistances, is so high that it interferes with proper circuit operation. The current in a circuit is normally $V/(R_s + R_L + R_m)$, where R_s is the internal resistance of the power supply and R_L is the load resistance. If the meter resistance (R_m) is a significant fraction of the other two resistances, then the meter will read less current than actually flowed in the circuit before the meter was installed.

Obtaining high DC current scales

The basic DC meter movement has a single current scale like 0 to 1 mA, 0 to 100 μA, and so forth. You can measure larger currents if you connect a shunt resistor in parallel with the basic meter movement (see Figure 4). In some cases the meter shunt is internal to the meter movement (Figure 5A); in other cases it's external (Figure 5B). An external shunt is usually bolted to the meter terminals (Figure 5C).

For shunted meters, the actual meter movement full scale rating is often printed in small letters on the lower right or left side of the scale. By the way, it's common for a meter that normally takes an external shunt to show up at a hamfest *sans shunt*. You might think that the meter is rated for 0 lo 1000 mA, only to find out that it is a 0 to 1-mA movement and requires a shunt to make it read the higher scale! Watch out; this can destroy the instrument.

The full scale current (I_{fs}) flowing in the circuit of Figure 4 is given by Kirchoff's current law:

$$I_{fs} = I_m + I_s \qquad (1)$$

where

I_{fs} is the full scale current.

I_m is the current flowing in the meter coil.

I_s is the current flowing in the meter shunt resistor.

There are two basic methods for calculating the value of the shunt resistor — Ohm's law and the current divider equation.

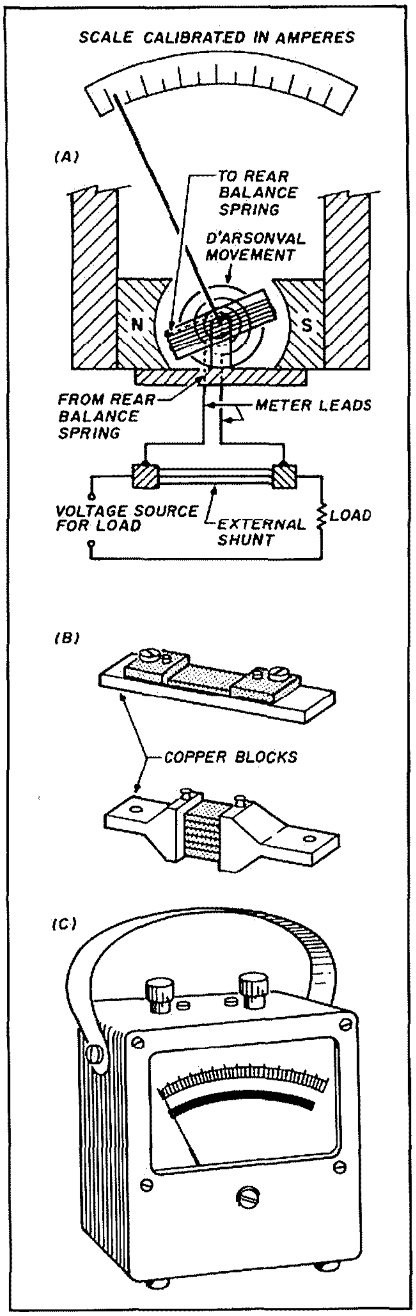
Ohm's law method

If you know the full scale current rating of the meter movement (M1) and the coil resistance (R_m), you know by Ohm's law that the voltage drop across the meter at full scale is $V_m = I_{fs}R_m$, or in the case shown in Figure 4:

$$V_m = I_{fs}R_m \qquad (2)$$
$$= 10_{-4}\ A\ (500\ ohms)$$
$$= 0.05\ volt$$

Because R_s is in parallel with M1, the same voltage is also present across R_s. Consider the case where you want the meter to indicate 1 mA (1000 μA) at full scale. Because 100 μA flows in the meter, you'd expect to find a current of I_s = 1000 μA $- I_{fs}$ = 900 μA in the shunt resistor when using Equation 1. You can therefore calculate the resistance needed to make a new full scale

FIGURE 5



(A) Internal shunt. (B) External shunts. (C) External view of meter.

current I_{fs}' by:

$$R_s = \frac{V_m}{I_s} \qquad (3)$$

$= (0.05 \text{ volts DC})/(9 \times 10^{-4} \text{ A})$
$= 55.556 \text{ ohms}$

Put in another form, this procedure reduces to:

$$R_s = \frac{I_m R_m}{I_{fs}' - I_{fs}} \qquad (4)$$

Current divider method

The alternative method takes advantage of the current divider equation:

$$I_s = \frac{I_{fs}' R_m}{R_m + R_s} \qquad (5)$$

If you solve this equation for R_s, you find the resistance of the shunt:

$$R_s = \frac{R_m (I_{fs}' - I_s)}{I_s} \qquad (6)$$

Voltage measurement from the DC current meter

You can measure voltage on a DC current meter if you connect a multiplier resistor (R_{mx}) in series with the meter movement, as in **Figure 6A**. The circuit is redrawn in **Figure 6B** to make it easier to understand. The current in the circuit is:

$$I = \frac{V}{R_{mx} + R_m} \qquad (7)$$

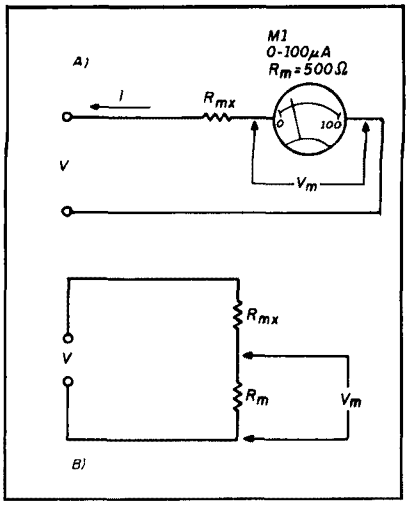
By solving **Equation 7** for R_{mx}, you can calculate the required value of multiplier resistor:

$$R_{mx} = \frac{V_{fs} - I_m R_m}{I_m} \qquad (8)$$

where

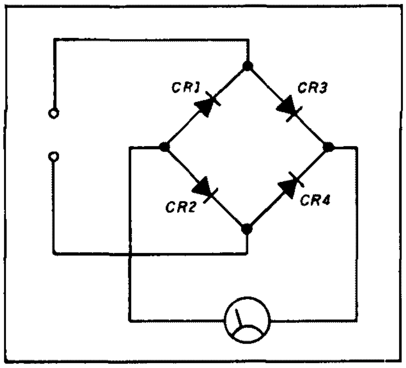
R_{mx} is the multiplier resistance.
V_{fs} is the desired full scale voltage.
I_m is the meter current at full scale.
R_m is the meter resistance.

FIGURE 6



(A) Multiplier resistor makes current meter into a voltmeter. (B) Circuit redrawn.

FIGURE 7



Bridge rectifier makes DC meter into an AC meter.

Voltmeter sensitivity

The sensitivity of the DC voltmeter is measured in terms of ohms per volt. It depends on the full scale current of the basic DC meter movement used in the circuit. You can find the sensitivity quickly by taking the reciprocal of the full scale current rating:

$$S = 1/I_f \qquad (9)$$

Table 1 gives the sensitivities of popular meter movements. The impedance of the DC voltmeter is found from its sensitivity:

$$Z = V_{fs} \times S \qquad (10)$$

You need to know the impedance when determining whether or not a particular meter will load the circuit.

Reading AC on DC meters

Although there are several different forms of AC meters, it's common practice to use a DC meter movement for measuring AC values in multimeters and other applications. The simplest method is to use a bridge instrumentation rectifier (see **Figure 7**). This rectifier circuit will produce a reading that's approximately 0.9 times the peak voltage, but you can calibrate the scale in terms of rms voltage. However, that calibration is based on the premise that the AC remains a sine wave. If the sine wave is distorted, or if a non-sinusoidal wave is being measured, then the calibration is shot.

The rectifiers used in **Figure 7** (CR1 to CR4) are, ideally, copper oxide instrumentation rectifiers for 60-Hz work. *You can use silicon (1N4148) or* germanium (1N60) rectifiers as well — especially if you're making RF measurements.

A more complex form of rectifier is based on the fact that op amp circuitry can be used to calculate the true rms value of the waveform. Several IC manufacturers offer devices that are listed as rms-to-DC converters.

Next month

This month I've examined the meter movement itself and some applications, like voltmeter and AC meter. In the next (and final) installment of this series I'll take a look at some meter applications circuits.

TABLE 1

Full Scale Meter Current, μA	Sensitivity, ohms/volt
1000	1,000
100	10,000
50	20,000
20	50,000
10	100 k

Commander HF-2500 Linear Amplifier

Command Technologies introduces the Commander HF-2500 linear amplifier. The HF-2500 operates on 160, 80, 40, 20, and 15 meters. It also performs on the WARC bands, and can operate on 12 and 10 meters with user modification.

This amplifier delivers 1,500 watts of continuous carrier output using a pair of Eimac 3CX800A7 triodes requiring only 50 to 80 watts of drive. High duty cycle emissions like RTTY, SSTV, FM, and AM present no problems. Tuned input and ALC adjustments can be made from the rear panel. The interior chassis layout provides full cabinet pressurized forced air cooling. Parts and labor are covered by a three-year limited warranty. With sales and service factory direct, the Commander HF-2500 sells for $2,188.

Write to Command Technologies, Inc., PO Box 939, Bryan, Ohio 43506 for further information or call toll free (800)736-0443.

Circle #303 on Reader Service Card.

CM-1200 SWR/Power Meter

NCG has announced the new CM-1200 Ultracompact SWR/Power Meter for 1200 MHz. It's rated at maximum power 60 watts for power ranges of 10 to 60 watts. Insertion loss is 0.25 dB and normal impedance is 50 ohms. The connectors are N type; dimensions are 2.25" width × 2.55" height × 1.1" diameter.

For more information contact NCG Company, 1275 N. Grove Street, Anaheim, California 92806.

Circle #304 on Reader Service Card.

Morse Code Terminal Kit and New Repeater Log

Dynamic Electronics Inc. announces a Morse Code Terminal Kit that converts a Radio Shack color computer into a Morse terminal. The conversion kit lets you send Morse code by pressing a key on the keyboard — decoding and printing the characters on the television screen or monitor. Interface cables (included) connect from the receiver's audio to the right joystick port and from the key jack to the printer or RS-232 port. The keyer cable is wired for a three conductor 1/4-inch plug and can be adapted to a two conductor 1/8-inch plug with an adapter (not included).

Features include automatic speed tracking on receive, preprogrammed message transmission with a single keystroke, and message transmission from a file on a tape or disk.

The program works on all 32K and larger Radio Shack color computers with extended BASIC using either tape or disk drives. The keyer will key a solid-state transceiver with a maximum keyup voltage of 15 volts. The cost of the kit is only $39.95 plus $3 shipping. Specify tape or disk software

Dynamic Electronics Inc. also has a new Repeater Log Program for Tandy color disk and IBM compatibile computers. The log lets you enter up to 200 repeaters in a file and list them by frequency, city, or callsign. Each file has spaces for city, repeater callsign, offset used, and comments. Additional files can be loaded for different beam headings or states. The cost is $14.95 plus $2 shipping. Please state IBM or Tandy color computer.

More information about the Morse Terminal Kit and Repeater Log Program can be obtained by contacting Dynamic Electronics Inc., PO Box 896, Hartselle, Alabama 35640. Phone: (205)773-2758.

Circle #305 on Reader Service Card.

Communicator Enhancer

Electronic Specialists improved modular connector communication enhancers now include Common Mode RFI and low frequency interference filtering. Available as integral parts of Spike/Surge Suppressors, the improved models provide modem and FAX transmissions with greatly reduced errors.

Dubbed the TURBO, improved communication enhancers are available for the standard single line modular plug (RJ-11) systems and commercial multi-line connector systems such as RJ-13, -14 and -45. Multiple connector systems can also be custom engineered. Models incorporating AC power line conditioners, filters or filter/suppressors are available.

TURBO models are available from stock, starting at $49. For more details contact Electronic Specialists, Inc., 171 So. Main Street, Natick, Massachusetts 01760. Call toll free: (800) 225-4876.

Circle #306 on Reader Service Card.

Two New Rotatable Dipoles

SV Products has two new rotatable dipoles. The model 1824D (25-foot elements) covers the two new WARC bands, 12 and 17 meters. The 2128D (22.4-foot elements) is for the DXer's bands, 10 and 15 meters. KD9SV uses a new trap design for high strength and power handling capability. The trap has been tested to 25 kV and will easily handle full legal power. Antenna elements are made from high strength T-60612 aluminum alloy. Windloading is less than 1 square foot and the antenna will rotate with a standard TV rotator. Introductory price for both antennas is $89.95 plus $5 shipping and handling.

For more information contact Gary Nichols, KD9SV, 4100 Fahlsing Road, Woodburn, Illinois 46797.

Circle #307 on Reader Service Card.

Transistor Radio Kit and Training Course

The AM-FM-108 Transistor Radio Kit is designed to expand your understanding of basic radio theory. Circuits are laid out in systematic order on an oversized pc board so you can easily understand the flow of the radio signals from antenna to speaker.

Construction begins with the building of the AM Radio, static testing of each circuit, and then dynamic testing with voltage and signals. After completing the AM Radio section, you construct and test the FM Radio circuit. Alignment procedures complete the projects. The radio kit pc board needs no cabinet. A special bracket provides support to use or display the radio in any location.

A complete manual is included along with a magic wand that assists in the alignment of RF circuits. Earphone, battery, and magic wand are included. The list price of the kit is $29.95.

For further information contact Elenco Electronics, Inc., 150 W. Carpenter Avenue, Wheeling, Illinois 60090. Telephone: (312)541-3800. FAX:(312)520-0085.

Circle #308 on Reader Service Card.

Scanner and Shortwave Answer Book

Bob Grove's *Scanner and Shortwave Answer Book* answers hundreds of questions asked by shortwave listeners. This 160-page reference was compiled from eight years of questions submitted to *Monitoring Times* by its readers.

Grove's book is $12.95 (plus $2 shipping in the U.S.) and is available from Grove Enterprises, PO Box 98, Brasstown, North Carolina 28902. Telephone (704)837-9200.

Circle #309 on Reader Service Card.

ICOM Options for the IC-901 Fiber Optic Multi-band Transceiver

The IC-901 fiber optic remote mount multiband transceiver offers a variety of options to meet your mobile needs. New options available to complete your IC-901 multiband package include:

Unit	Description	Suggested Retail Price
UX-S92A	2-meter SSB band unit complete with USB, LSB and CW modes	$599.00
SP-12	Slim Line External Speaker with Visor Clip for mobiles	$ 25.99
OPC-23	DC cable	$ 33.49
EX-766	Interface-A Connector	$ 47.99
EX-767	Interface-B Connector	$ 42.99

For details contact ICOM America Inc., 2380 116th Avenue NE, PO Box C-90029, Bellevue, Washington 98009-9029.

Circle #310 on Reader Service Card.

Kantronics Data/ Voice Radio

Kantronics has introduced the first radio specifically designed for the packet enthusiast. The new Kantronics dvr 2-2 provides high speed T/R switching for today's digital world.

The dvr 2-2 reaches full power output (rated at 2 watts) within 5 ms after push-to-talk is activated. In addition, the dvr 2-2 has a fast carrier detect output that reacts within 10 ms after a signal is received. This lets you set your TXDELAY to 2 (20 ms) when talking to another dvr 2-2 user. This carrier detect helps reduce collisions, as the radio senses activity faster and signals the TNC that the channel is in use.

The dvr 2-2 also has discriminator output available on the rear panel connector, which means you don't have to modify the radio for high speed packet. The unit was designed with the ability to operate at up to 9600 baud without modification. The connections to your packet modem are all provided on a single DB-9 data port on the rear panel.

The dvr 2-2 data port is designed to be pin-for-pin compatible with Kantronics TNCs, but the dvr 2-2 will operate with other manufacturers' units also. The increase in speed realized by our pin diode T/R switching can result in a reduction of your TXDELAY setting by as much as 90 percent or more.

The dvr 2-2 has a microphone jack on the rear panel, allowing you to connect the optional Kantronics mic and an external speaker for voice operation. You don't have to disconnect your packet unit to talk.

The suggested retail price for the dvr 2-2 is $199. For details contact Kantronics, 1202 E. 23rd Street, Lawrence, Kansas 66046. Telephone: (913)842-7745.

Circle #311 on Reader Service Card.

Heil Sound's Concept 2000

Heil Sound has introduced the Concept 2000 product line. Included in the lineup are: the HM-10 microphone, AB-1 adjustable microphone mount, FS-1 foot switch, and FL-1 private lighting system.

The Heil HM-10 uses the Heil "Key Element" microphone cartridge for maximum articulation of SSB transmissions. The HM-10 is available with either the HC-5 or the HC-4 Key Element. The HC-5 is a full range element rolling off below 300 Hz and has a 6-dB peak at 2100 Hz. The HC-4 "DX Dream Machine" was designed by Heil for breaking the DX pileups. It has the last octave rolled off at 600 Hz and has a 10-dB peak at 2100 Hz.

A special model of the HM-10 is available with *both* the HC-4 and the HC-5 installed. A professional four-pin Cannon connector in the base of the HM-10 is used to connect the microphone to your transceiver. A second micro-switch selects either the HC-4 or the HC-5.

The HM-10, shipped with mating cable assemblies, will match the HM-10 with your type of transceiver. The red cable is wired for Kenwood. Blue mates the ICOM, yellow matches up the YAESU, and a white cable is used for other transceivers and labeled as such. The slide switch on the HM-10 case is wired for standard push-to-talk while the Key Element is wired straight through for VOX operation all the time without special switching or cabling.

The AB-1 is an adjustable mount to support the HM-10 or similar professional microphones. Its adjustable boom and mounting assembly can be used in a variety of ways.

The FS-1 has been designed to control the push-to-talk lines of a radio transmitter, allowing hands-free operation. This device has a 2-A momentary switch connected to a 4-foot shielded cable and is designed to switch low voltage circuitry. Switching 110-volt AC circuits with the FS-1 is not recommended.

The cable is terminated with a 1/4-inch phone plug. Should your equipment require a different interface, it is not recommended that you use any type of adapter but rather replace the existing connector with the correct matching plug.

The FL-1 is a private lighting system to illuminate radio panels, log sheets, or operating positions. The FL-1 is shipped with a conventional 12-volt incandescent lamp that can be powered with any 12 to 14-volt 100-mA power supplies. A halogen bulb is available to fit inside the FL-1 should you want more candlepower.

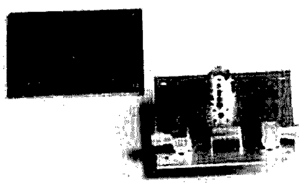
Additional information on Concept 2000 products is available from Heil Sound, Ltd., 2 Heil Drive, Marissa, Illinois 62257. Telephone: (618)295-3000.

Circle #312 on Reader Service Card.

Evaluation/Prototyping Board with Sockets: TFM Series Mixers

RF Prototype Systems introduces the TFM1 Quick Board. The TFM is for evaluation or breadboarding using the following Mini-Circuits devices, and equivalent devices manufactured by Pulsar Microwave, Tele-Tech, Engelmann Microwave, Synergy Microwave and Olektron: TFM mixers, TSC splitters and TDC couplers. Connections to BNC, SMA, or SMB connectors are via 50-ohm microstrip lines.

Mounting holes for three bulkhead BNC, PCB, SMA or SMB connectors are provided.



A socketed version, TFM1S, allows the devices and TFM1S to be reused easily without soldering for frequencies up to 2 GHz.

This board is double-sided FR-4 (high temperature G-10) with plated-through holes.

For more information, contact RF Prototype Systems, 12730 Kestrel Street, San Diego, California 92129. Phone: (619)538-6771.

Circle #313 on Reader Service Card.

UNDERSTANDING COMPUTER-GENERATED RFI

Some remedies for this malady

By Bryan P. Bergeron, NU1N, 30 Gardner Road, Apt. 1G, Brookline, Massachusetts 02146

Radio Frequency Interference (RFI) has always been a source of concern for the Radio Amateur. Until the recent introduction of complex circuitry in the home (like microcomputer systems, VCRs, and microwave ovens), our communications gear has been the main source of potential RFI. In this computerized age of ours, increasingly complex and sensitive receivers are being barraged by these and other sources of RFI.

Microcomputers have become an integral part of the contemporary Amateur Radio station. Virtually all modern receivers and transceivers rely on microcomputer-controlled circuitry for their internal operation. But more important, from the perspective of RFI generation, is the external standalone microcomputer system. This system can be found in an ever-increasing number of Amateur Radio stations running programs for predicting HF propagation, logging and checking QSOs, printing QSL cards, and even controlling transceivers. Unfortunately, many computer systems radiate a significant amount of RF into the shack. This is especially true on the HF bands, where even a small amount of RFI can mask an otherwise readable signal. This article examines the microcomputer as a source of RFI, and suggests some steps you can take to contain it.

To understand how the microcomputer can be a source of RF interference, you must have some knowledge of how microcomputers are constructed, how they operate, and how they are normally connected to other devices.

Physical Construction

On the most basic or physical level, most microcomputers are composed of a system unit, power supply, keyboard, and display screen. Depending on the make and model of the microcomputer, these components may be physically separate, as in the IBM PC, or found in a single package, as in many portable computers. Many microcomputers include the power supply and at least one floppy disk drive in the same physical enclosure as the system unit. Other models have the power supply and disk drives packaged separately. Many of the popular microcomputers have slots for extending the basic system with plug-in modems, memory cards, and video cards for a variety of screens.

Few of us make do with the minimal system configuration. The most common additions are printers, extra disk drives, and modems. Adventuresome hams have packet controllers and interfaces to their microcomputer-based transceivers. All of these additional devices, and the cables that connect them to the system unit, are potential sources of RFI.

Logical operation

For the purposes of this discussion, consider the microcomputer to be composed of a Central Processing Unit (CPU), memory, and a system clock. The CPU is the heart of the microcomputer. It not only performs operations on data that resides in memory, but also keeps track of the current status of the executing program and handles communications with memory and input/output devices like printers and modems. All of this activity must be carefully orchestrated for the microcomputer to function properly. The metronome for this activity, and a potential source of RFI, is the all-important *system clock*.

The system clock creates the timing signals used to synchronize all activities within the computer. And as in our modern transceivers, a quartz crystal normally serves as the basis for this timing. In some microcomputer systems the CPU contains the oscillator circuitry, so an external crystal is simply connected between two pins of the CPU chip. In other systems, separate dedicated chips are used to generate the timing signals. In the IBM PC/AT, for example, there's a clock generator chip and a programmable timer chip. The clock generator chip, which uses a quartz crystal, creates the basic timing signals used by the computer

The programmable timer chip is related to the clock generator chip. Think of it as a programmable array of flip-flops that produces an output signal every so many clock cycles. For instance, if the basic clock cycle is 6 MHz (as in the original IBM-AT), and you want to perform some event every 1/60,000 second, the programmable timer chip can

TABLE 1

Clock frequencies for common microcomputer systems. Various speed-up methods and accelerator cards are used to speed up many of these machines (particularly older models).

Apple IIE/C	1 or 4 MHz
Apple IIgs	1 or 6 MHz
Commodore 64	1 MHz
Compaq Deskpro-286	8 MHz
IBM AT	6 (original) and 8 MHz (later)
IBM PC	4.77 MHz
IBM PS/2 Model 30	8 MHz
IBM PS/2 Model 50	10 MHz
Macintosh 512/Plus/SE	7.8 MHz
Macintosh SE-30	16 MHz
Macintosh II, IIx, IIcx	16 MHz

be programmed with a count of 100. At every 100th clock cycle, the programmable timer chip will produce a signal that can be used by the computer circuitry.

Obviously, RF energy at both the clock frequency and the variable programmable frequency (6 MHz and 60 kHz in the example above) are of concern to the Amateur. See Table 1 for a listing of clock times used in the popular microcomputer systems. In some cases, the basic crystal oscillator circuitry operates at the same frequency as the system clock. For example, the 10-MHz IBM PS/2 Model 50 uses a 10-MHz crystal. In other instances, a crystal oscillator with a frequency higher than the clock frequency is used in conjunction with frequency divider circuitry. For example, the IBM PC uses a 14.32-MHz crystal with a clock frequency of 4.77 MHz (14.32 divided by 3).

Power

Most microcomputers with built-in power supplies, like the Apple Macintosh and IBM PC/AT and PS/2 series, make use of internal lightweight switching power supplies. Switching supplies, unlike conventional linear supplies, do not make use of large iron core power transformers. Instead, the AC from the 110-volt power line is directed to a bridge rectifier, and the resulting ripple DC is pulsed at between 20 and 100 kHz. This pulsed, high frequency DC allows for the use of small, lightweight high frequency transformers. The pulsed DC and its harmonics are potential sources of RFI, through the power lines and cables to computer accessories. Also, whereas the relatively massive, high inductance power transformers effectively block RF radiation into the power lines, the high frequency transformers used in switching supplies can potentially couple RF more easily into the AC power line.

Peripherals

Sooner or later most of us will add an electronic keyer, power amplifier, or beam antenna to our bare-bones rig. Similarly, those of us bitten by the computer bug are seldom satisfied with a minimally configured microcomputer. The most common additions include floppy and hard disk drives, various types of printers, modems, and alternative input devices. Depending on your computer's design, many of these additions may take the form of cards that plug into the system, or external peripherals that must be connected by cables to the system unit. In order to understand how these devices can cause RFI, you must have a basic grasp of their operation.

Alternative input devices

Although the keyboard is by far the most common method of interacting with the computer, there are various alternative input devices. What follows is a short description of the most common ones.

The Mouse: Some micros, like the Apple Macintosh, come factory equipped with this cursor control device. Other systems, like the IBM PS/2 series, have a mouse interface which lets you select a mouse of your choice. From an RFI perspective, the mouse, which is the most popular alternative or supplement to keyboard entry,[1] can be classified as mechanical or optical. The mechanical mouse uses a roller ball that moves as you control the mouse. Two perpendicular rollers (one for the x-axis and one for the y-axis) attached to contact pins are coupled to the roller ball. These contact pins make and break connections with a contact bar as the mouse moves, much like a distributor in an automobile engine. Optical mice are similar to mechanical mice in many respects; however, there are no physical make-and-break connections. They use LEDs and phototransistors to detect motion. As you might expect, the rapid make-and-break connections associated with mechanical mice can result in RFI. In comparison, optical mice are more electronically "clean." But, like any peripheral attached to the system via a cable, the mouse cord can act like a broadcast antenna for signals inside the system unit.

Trackballs: Trackballs are best thought of as inverted mechanical or optical mice. They offer the same benefits and limitations as the other mice in terms of their potential for RFI.

Light Pens: Light pens work by sensing the exact time that the electron beam in the monitor excites a phosphor at a particular point on the screen. The associated circuitry of the light pen determines the x-y coordinates of the point on the screen by measuring the time it takes for the electron beam to reach the pen. Light pens pass this information back to the computer through a cable, or in some cases, by sending an RF signal to a receiver mounted on the top of your monitor.[2] In directly wired systems, light pens generate low intensity signals only every 1/60 of a second. Assuming that the cable is adequately shielded, the potential for RFI is relatively low. RF light pens, in comparison, have a high RFI potential.

Tablets: Graphic tablets, useful for drawing and tracing, come in three basic types: electromagnetic, resistive, and acoustic.[3] The most RFI-prone type, the electromagnetic version, has a handheld pen that transmits an RF signal to a receiving grid located under the tablet surface. Tablet circuitry converts the signals into x-y coordinates to determine the pen's exact location.

Resistive tablets, sometimes called touch pad tablets, are made of two conductive surfaces separated by a small air gap. When a pen touches the tablet, bringing the two surfaces together, current flows between the two surfaces. The strength of the current is used to determine the x-y coordinates of the pen.

Acoustic tablets use a pen transmitting ultrasonic waves (65 to 75 kHz) that are received by microphones near the

work area. Through triangulation, the relative strength of the received signal at each microphone is used to calculate the relative x-y coordinates of the pen.

Touch screens: Touch screens, like light pens, are useful for selecting objects on the screen. The most common varieties are mechanical, optical, and capacitive. Optical screens use rows and columns of infrared LEDs — phototransistor pairs mounted opposite each other along the edges of the screen. Touching a particular point on the screen with your finger blocks one or more x-y beams of light. The touch screen sends the coordinates of the broken beams to the computer, which calculates the corresponding x-y location on the screen.

Mechanical switching panels are composed of transparent, conductive membrane switches mounted over the display screen. Pressing your finger on the panel brings the two conductive surfaces together and completes the circuit. One sheet determines the x-axis and the other the y-axis location of the contact.

Capacitive touch screens have a capacitive coating on the CRT screen that acts as one plate of a capacitor. When you make contact with the screen, current flows into your body from the contact point. Sensors on the screen detect the location of the current drain and calculate the corresponding x-y location. In my experience, capacitive systems are more prone to RFI than either the mechanical or optical versions.

Joysticks: Although they are more popular as a game interface than a way of manipulating Amateur software programs, joysticks should not be overlooked as a source of RFI. The vast majority of joysticks are mechanical, composed of switches and/or potentiometers. The largest RFI threat from these simple devices occurs when the connecting cable acts as a radiator for system unit signals.

Modems

Modems (named for **MOD**ulator-**DEM**odulator) let digital computers communicate over analog phone lines. In some microcomputer systems, like the Apple II series and the IBM PC/XT/AT, modem cards can be inserted easily into slots in the system unit. Along with minimizing the potential for RFI, these internal modems have the added benefit of providing a less cluttered ham shack. The more RFI-prone external modems, also popular on the IBM PC and other microcomputers, must be connected by a cable to the system unit. Purchasing an internal modem doesn't guarantee freedom from RFI, however, because the telephone cable represents a potential RF antenna.

Printers

Printers range from simple dot-matrix units to complex laser printers that contain their own RF-producing microcomputer systems. Although a few printers attach directly to the system unit (primarily on portable models), the vast majority are connected to the system via cables. In my experience, the mechanical printers are less likely to cause RFI, but the acoustic noise they produce is hardly bearable during a QSO. The relatively silent laser printers, by comparison, emit considerable RF energy.

RF modulators

Many of the lower priced microcomputers are designed to work with TV receivers as monitors. RF modulators convert the video signal into a VHF signal that can be handled by a TV receiver (commonly on channel 3 or 4). These so-called "RF bricks" are potential sources of RFI — especially when the output is unshielded 300-ohm flat line.

Local area networks

A local area network lets microcomputers communicate with other devices connected to the network, including modems, printers, computerized communications gear, and other microcomputers. ICOM's system allows multiple ICOM receivers and transceivers to communicate with each other and with microcomputers. Because this system directly connects your computer system to your communications gear, there's ample opportunity for RFI. In my experience with the ICOM system, there's no detectable RFI as long as the integrity of the cable and associated connectors are maintained. Other networks, like Apple's AppleTalk, can cause considerable RFI. AppleTalk is a low speed network often run over standard (unshielded) telephone cable. Receive on my ICOM 751A is rendered practically useless when AppleTalk is in operation.

Minimizing computer-generated RFI

There are a number of steps you can take to minimize computer-generated RFI in your shack. In some cases, your problem may be cleared up by following only one or two of these measures. In more difficult situations, you may have to try all of these suggestions for acceptable results.

Check the FCC rating before you buy: Microcomputers are rated by the FCC as either Class A or Class B devices, depending on the amount of RFI produced by the equipment (see **Table 2**). Paradoxically, the often cheaper Class B machines, intended for home use, are less prone to RFI than the Class A machines. The more expensive Class A micros, including many micros based on the Intel 80386 chip and several of the large monitors, have less stringent RFI ratings. Unfortunately, many hams use the more powerful and more RFI-prone "business class" machines at home.

Think twice about a micro with a Class A FCC rating. Taming a machine with a Class B rating is difficult enough! Check the Model/Serial number tag on all computer equipment, including modems, printers, scanners, and mice before you buy. You should see an FCC ID number, together with a statement like "Certified to comply with the limits for Class B computing device pursuant to Subpart J of Part 15

TABLE 2

Radiation limits applied to microcomputers manufactured after January 1, 1981, based on data from subpart J of Part 15 of FCC rules.

Maximum radiation measured at 3 meters	Class A	Class B
30-88 MHz	3,000 μV/m	100 μV/m
88-216 MHz	5,000 μV/m	150 μV/m
216-1000 MHz	7,000 μV/m	200 μV/m
Maximum conduction into the AC power line		
0.45-1.6 MHz	1000 μV	250 μV
1.6-30 MHz	3000 μV	250 μV

FCC rules." These rules are designed to provide "reasonable" protection against RFI in a residential installation.

Shielding: Don't defeat the shielding in your computer or peripherals. You may be tempted to remove the aluminized plastic backing from the Macintosh SE or Commodore 64 motherboard to prevent heat buildup. Don't! You will have a very cool-running RFI machine. Also, keep the rear metal card covers on your system unit intact. If you remove an internal card, make certain that you replace the original slot cover.

The new IBM PS/2 machines, like the original Macintosh series of computers, make heavy use of metalized plastic for shielding. Unlike the original PC and many of the PC clones these new lightweight machines limit metal shielding mainly to the power supply. If you aren't careful when opening these plastic cases, you might chip or wear away the conductive paint, and have a less than perfect RFI shield.

Cables: Use shielded cables whenever possible, and add snap-on ferrite inductors to peripheral cables — especially if they aren't shielded. You can use ferrite snap-on toroids to increase the series inductance of cables, raising their impedance to HF signals. Although adding significant inductance may have the effect of reducing the computer signals, the high frequency RF components will be attenuated to a greater degree. Peripherals (like external disk drives) that can create their own RF signals should have ferrite snap-on inductors attached to both ends of the cable. The inductor near the peripheral attenuates signals generated from within the peripheral, while the inductor near the computer system attenuates signals generated by the system clock that may be inadvertently coupled to the peripheral "antenna." Don't forget to add an inductor to the telephone cable where it exits your modem.

Bypassing: Judicious use of RF bypass capacitors with resistive touch pads, mechanical mice, and joysticks often pays off.

Power Conditioning: The simplest way to provide a good degree of isolation between your computer equipment and your receiver is to make certain that each system is controlled by different circuit breakers. Plug your computer and peripherals into a wall socket that is not connected to the socket used for your communications gear.

If using separate power circuits fails to remedy your RFI problem, or if all of the sockets in your shack are controlled by a single circuit breaker, try adding two good surge protectors to your shack — one for your communications gear and one for the computer equipment. A simple protector with MOVs won't do. The best method of isolating signals coupled through the power line uses a combination of RF line filters and transient suppressors. You can realize a 60-dB attenuation of interference above a few hundred kHz with RF line filters.[4]

Ground: Although you have no doubt heard it before, a good ground is essential for minimizing RFI. It's surprising how many hams who have 6-foot ground rods connected by heavy coaxial braid to their gear fail to ground their computer equipment. Treat your computer system, including all peripherals, like your communications gear where ground is concerned, and you should be well on your way to minimizing potential RFI.

Layout: Minimize cable lengths. When possible, use an internal modem instead of an external one with its associated cables. If RFI persists, try rearranging your equipment. Move your micro and peripherals as far from your receiver as possible. In some cases, interference can be minimized to acceptable levels through proper layout of equipment and judicious cable runs. Obviously, running your external disk drive cable parallel and adjacent to the antenna feedline is asking for trouble.

Communications Gear Modifications: Try to minimize the number of possible entry points for computer-generated RFI into your system. If you have an external speaker with more than a few inches of cable, use a low pass filter and shielding to prevent the speaker wire from acting as an antenna.

Software Design: If you develop your own software, try to minimize the reading and writing of data to disk. Similarly, when you purchase software developed by others, run the program and make note of how often the disk drive whirs. The stepper motors and associated drive circuitry are extremely noisy in the RF spectrum.

Summary

The best way to handle computer-generated RFI is to think of your computer system as you would any other piece of RF communications gear, with peripheral cables, phone connections, and power cords acting as the antenna system. Use low pass filters on all antennas (snap-on toroids on all cables and power cords), make certain that you provide a good system ground, use shielded cables of minimum length, and start with "RF-clean" gear. Use bypass capacitors whenever possible, and keep the computer "antenna system" away from your communications gear. **hr**

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3. A. Poor, "Tablets for Precision Graphics," *PC Magazine*, 1987, pages 159-166
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AN LED MILLIVOLTMETER

By Yardley Beers, WØJF, 740 Willowbrook Road, Boulder, Colorado 80302

I've found a way to use light-emitting diodes (LEDs) to display rapid changes in voltage. Here are the construction details for a device with four LEDs where the number illuminated depends upon the input voltage. Without amplification, the voltages which turn on the lights are about 100, 200, 300, and 400 mV, respectively. A built-in DC amplifier can be switched in to increase the voltage sensitivity by a factor of 100.

Some uses

A gadget of this type can serve as an inexpensive and compact substitute for a cathode-ray oscilloscope. You can display changes in voltage and see them from a considerable distance. It's especially useful when the voltages are changing very rapidly. You can also use it to display the output of a signal strength meter which is remotely located from a circuit being adjusted for maximum signal strength. The device is useful for demonstrating the generation of induced emfs to a class of beginners. Just connect a 1-1/2 inch coil with 25 turns to the input. Using a string, pull a small bar magnet through the coil. All four LEDs will light briefly.

I've included information for designing your own instrument using different numbers of LEDs to improve the resolution. You can also make the response logarithmic instead of linear.

A recent *Ham Radio* article discussed many of the fundamental properties of LEDs, but didn't mention their fast response times.[1] In fact, it's possible to build an inexpensive but low resolution oscilloscope using a square array of LEDs.[2]

How it works

The principle of operation is illustrated by the circuit in **Figure 1**. Its principal components are an LED and an integrated circuit, U1 — a comparator. This IC, like an op amp, has two inputs and an output. However, there's no internal feedback which tends to make the voltages at the input terminals equal. The LED and a current-limiting resistor, RL, are connected in series between the output terminal and the positive terminal of the battery, which powers the instrument. The signal voltage is applied to the negative input terminal, while the positive input terminal is connected to a voltage determined by resistive voltage divider R1 and R2, connected between ground and the battery's positive terminal. The comparator acts as a switch which turns on the LED when the input voltage exceeds a critical value. This value is determined by the voltage on the positive input terminal. Such a circuit serves as a building block for a more complex instrument made from a number of these units, with the negative inputs connected in parallel and the positive input terminals connected to various taps on a voltage divider.

The value of current-limiting resistance RL is found from Ohm's law by dividing the battery voltage minus the drop across the LED and comparator (about 2 volts for red and 3 volts for green LEDs) by the desired current.

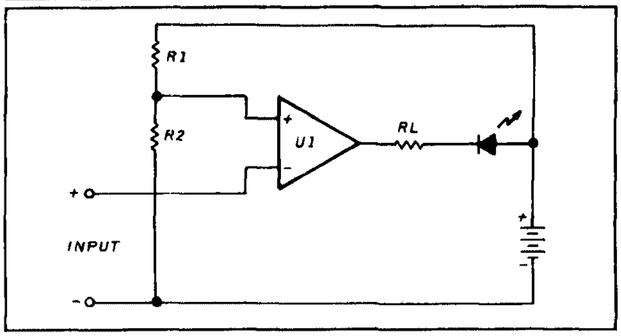
A practical instrument

A simple instrument based on **Figure 1** could have some uses in the Amateur station. However, ICs with multiple comparator circuits are available. Chips containing four units (type 339) are the most common and are available from Radio Shack and elsewhere. My device is based on one of these chips. The circuit in **Figure 2** is a composite of two previously published diagrams. The more important portion is taken from page 87 of **Reference 3**.

The voltage divider, which establishes the reference voltages of the positive terminals, consists of four 1-k resistors in series with a 100-k trimpot. The trimpot sets the input voltage at which the first LED turns on, normally about 100 mV. It's possible to make it turn on at about 40 mV but with reduced brilliance.

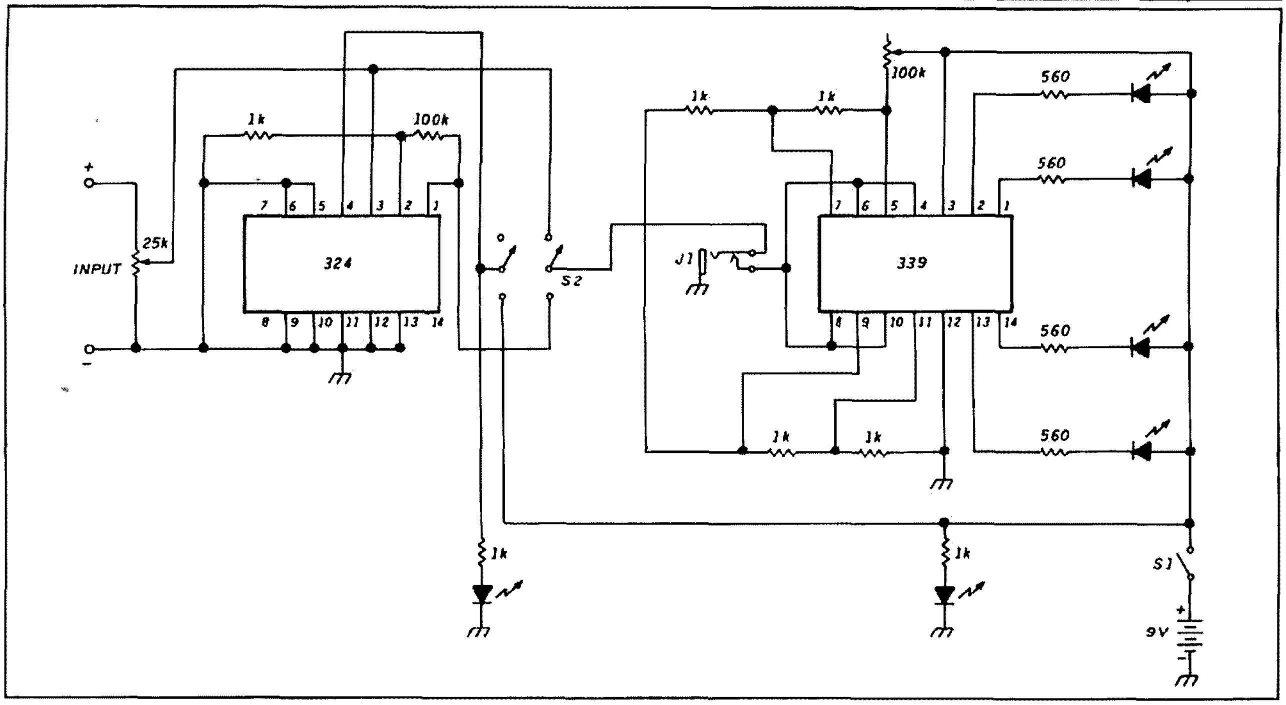
The portion of the circuit containing the 324 op amp is based on **Figure 3** of **Reference 4**, but the gain is held constant by making the feedback resistor fixed at 100 k. The gain control, consisting of a potentiometer across the input, can be used with the amplifier both in and out. I've included

FIGURE 1



Basic circuit diagram. When the input voltage exceeds a value determined by the voltage on the + terminal, comparator U1 acts as a switch and turns on the LED.

FIGURE 2



Circuit diagram of a practical unit. This uses the four basic units shown in Figure 1. Also included is a DC amplifier using one section of a 324 quad op amp that can be switched into the circuit.

PARTS LIST

Solid-state devices:
1 Type 339 quad comparator IC
1 Type 324 quad op amp IC (only one section used)
4 10-mA red LEDs (for display)
2 Miniature red lEDs (for pilot lights)
Resistors:
4 560 ohm, 1/4 watt
7 1 k, 1/4 watt
1 100 k, 1/4 watt
1 25-k potentiometer
1 100-k trimpot
Hardware:
1 Sardine can (small size)
2 Binding posts
1 SPST switch
1 DPDT switch
1 1/8" closed-circuit jack
1 Connector for 9-volt battery (salvaged from dead battery)

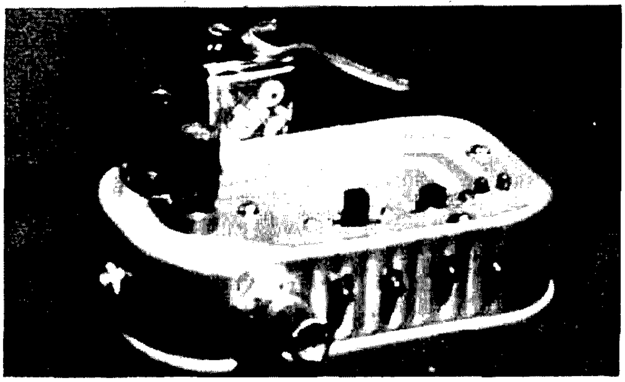
closed-circuit jack which makes it possible to use an external meter to measure amplifier input or output.

Construction

The device is contained in a small 4 × 2-5/8 × 1-inch sardine can with a 9-volt battery clamped to the rear 4 × 1-inch surface, as shown in Photo A. Binding posts for the input, two slide switches, and two miniature pilot light LEDs are located on the top. The gain control and four 10-mA red LEDs for the output display are on the front. On the left you'll find the 1/8-inch closed-circuit jack connected to the op amp output.

The ICs and resistors are mounted on a 3 × 1-1/4 inch piece of perfboard (see Photo B). The board is supported from the top surface of the can by three 4-40 machine screws with 1/4-inch spacers.

PHOTO A

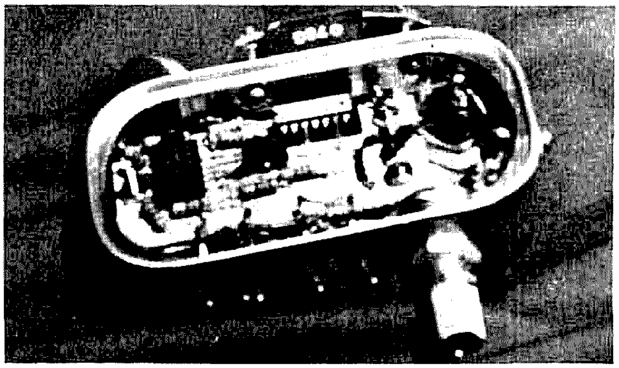


General view. The device is housed in a small sardine can. The four display LEDs and the gain control are on the front. Input binding posts, two switches, and two pilot-light LEDs are on the top.

I punched holes in the can with an awl, enlarging them with a drill or reamer when necessary. I mounted the LEDs by forcing them into tight fitting holes that I had carefully enlarged. I expected to use some household cement to hold them in, but I found this was unnecessary.

To mount the battery, I took two strips of sheet metal 2 inches × 1/2 inch and drilled matching pairs of clearance holes for 6-32 screws 1-3/8 inches apart. Using one of the strips as a template, I punched a matching pair of holes on the back surface of the can. Then, with one of the strips inside the can to strengthen it, I placed 1-inch bolts through

PHOTO B



Bottom view. The resistors and ICs are mounted on a piece of perfboard supported from the top of the can by three machine screws with spacers. The 324 op amp IC is to the left; the 339 comparator IC is at top center.

the holes facing outward. I then attached lock washers and nuts on the outside. The other strip and another pair of nuts hold the battery in place.

I wired the components on the can and perfboard separately as best I could. I provided terminals on the perfboard or leads that extended beyond it for the interconnection between the board and the can. I spliced some of these leads to those of the LEDs. Because it's easy to make connections the first time but difficult to separate them for correcting wiring errors, I checked the wiring on the perfboard very carefully with a volt-ohmeter as I went along. This care paid off; the device worked the first time I turned it on.

Discussion

Because decimal number systems are commonly used, you might like to have ten LEDs rather than four. You can build such a system with three type 339 chips, leaving two comparators unused. I think it's logical to arrange the voltage divider so the input-voltage increment which causes adjacent LEDs to light up is one-tenth rather than one-quarter of the voltage needed to light the last LED. This improves the resolution. Actually, ICs with ten comparators on the same chip are designed for this purpose. These chips are known as "bar graph drivers." A bar graph is a chip containing the equivalent of ten LEDs. These chips are often built into various instruments like signal strength meters. The bar graphs are more convenient in such applications, but are visible only at short distances.

There are other possible variations in the design of this device. You can arrange the taps on the voltage divider so the input-voltage increments correspond to a fixed number of decibels rather than a fixed amount of voltage. You can make a device with considerable flexibility by using a switch which connects the positive terminals of the comparators to different sets of taps on the voltage divider. Circuit diagrams for some of these alternative designs are found in Reference 3.

It's easy and fun to build this project. You can amuse visitors with demonstrations of induced emfs, antenna patterns, and a variety of physical effects.

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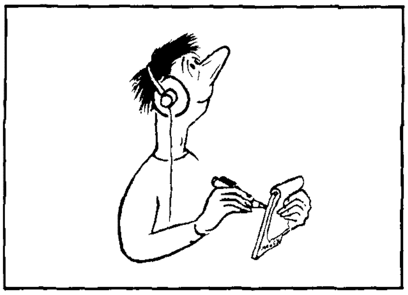
DX Forecaster

By Garth Stonehocker, KØRYW

SPRING FREQUENCY CHANGES



It's time for spring; winter is just about over. Are you ready for a change? How did your DXing fare this winter? Are your antennas still up? Is your operation in need of maintenance or changes? Why not get to work on these items before spring cleanup or summer yard work starts taking up your time?

Several aspects of propagation that affect DXing change during the spring and fall. The length of the day, while nearly equal to night (exactly equal this year on March 20th at 2119 UTC), changes rapidly from shorter to longer during this month and into the first weeks of April. This means that the maximum usable frequencies (MUF) for a particular path change quite rapidly. The HF commercial radio users — mainly maritime/aeronautical and broadcasters — transmitting point to point or to specific areas, need to obtain new frequency allocations from the International Telecommunication Union (ITU) during a two month interval. The annual allocation intervals of frequency usage are March to April, May to August, September to October, and November to February. Their allocations are in bands like ours: roughly 300 to 900 kHz wide at 2, 4, 6, 8, 12, 16, and 22 MHz for maritime radio, and 200 to 350 kHz wide at 6, 9, 11, 15, 18, and 21 MHz for broadcasters. Each of the maritime users have calling and working frequencies in each band. The ships use one set, the shore stations the other. They are able to use any band that's propagating to shore or ship stations in order to transact their business, so this operation mode automatically takes care of the seasonal changes of MUFs for the distance involved. Older operators have a working knowledge of propagation. To help the new operators, the shore stations send ship companies charts of frequencies to use out to concentric circles of distance at sea from the station. Operators listen to the stations calling or working the bands to verify which will work for them. This system serves them well even when using selective calling to a specific ship. Aeronautical traffic control and the airlines, through aeronautical communication companies like ARINC or Collins, use a similar system.

The broadcasters transmit in one or two bands of allocated frequencies that should work to a target area.* The frequency and time schedule is sent to those listeners who request it and is published in magazines like *Popular Communications*, *Monitoring Times*, or *World Radio TV Handbook*. Broadcasters are the main users of the ITU's four allocation intervals to change frequency with the season and sunspot cycle. They are restricted to this predetermined schedule, unlike the maritime/aeronautical operators who can jump from band to band until the message is passed. Those who listen to broadcast stations need to know they will be broadcasting where the published schedule indicates. Limited schedule changes (usually for QRM) are allowed by the ITU, but at the broadcasters' risk of losing their audience. You can imagine the chaos if all the broadcasters began shifting frequencies and bands at will! By using experienced propagationists, the broadcasters request frequency allocations. The ITU then uses its computers to juggle frequency assignments for the hours of the day for each season. It can be chaotic keeping signals within the bandwidth (adjacent channel interference), because propagation prediction is far from reality (by months, days, or hours) and some countries don't abide by the rules. Yet this system works pretty well, as far as propagation is concerned, for the seasons and sunspot cycle.

Hams have quite a bit of freedom to shift frequencies and bands at will. We are limited only by our knowledge of propagation and our equipment (antennas on hand) to change frequency with the hour of the day, season, and sunspot cycle to work the DX or get the message through. You can see why communicators put so much significance on propagation for frequency allocation. It's important for hams too!

Last-minute forcast

Conditions will be excellent for DX on the higher frequency bands (10 to 30 meters) the second and third full weeks of March. The MUFs will be high because of a solar flux maximum. Openings will be noticeably longer into the evening. Late evening one long hop transequatorial openings are most probable on the 5th, 14th, 23rd, and 31st, when a disturbance in the geomagnetic field and ionosphere is expected. The lower bands should provide the best DX around the 5th. Spring thunderstorms may cause local noise, but strong signals should overcome the noise — except for the actual time of the flash. Poor signals and QSB during the disturbed periods could be the big problems this month on the lower frequencies, especially on east-west paths. Spring equinox occurs on March 20th at 2119 UTC. A full moon appears on the 11th and will be at perigee on the 1st and 28th.

Band-by-band summary

Ten, 15, and 17 meters will be open from morning to early evening almost daily in most areas of the world. Expect higher band openings to be southerly, shorter, and closer to local noon. Transequatorial propagation on these bands is likely to be toward evening during times of high solar flux and disturbed geomatic field conditions.

Twenty and 30 meters will be useful almost 24 hours a day. Daytime conditions will resemble those on 17 meters, but skip and signal strength may decrease during midday on days with high solar flux values. Look for good nighttime use — except after

MARCH

GMT	PST	N	NE	E	SE	S	SW	W	NW
		ASIA FAR EAST	EUROPE	S. AFRICA	S. AMERICA	ANTARCTICA	NEW ZEALAND	OCEANIA AUSTRALIA	JAPAN
0000	4:00	15	30	12	10	10	10	10	10
0100	5:00	12	30	15	10	10	10*	10*	10
0200	6:00	12	30	15	10	10	10*	10*	12
0300	7:00	15	30	15	10	10	10	10	12
0400	8:00	17	30	17	12*	10	10	10	15
0500	9:00	20	30	17	12	12	10	10	20
0600	10:00	20	30	17*	12	15	10	10	20
0700	11:00	20	30	17	15	17	10	10	20
0800	12:00	30	30	17	15	17	12	12	20
0900	1:00	30	30	17	15	20	12	12	30
1000	2:00	30	30	17	17	20	15	15	30
1100	3:00	30	30	17	17	20	15	15	30
1200	4:00	30	17	12	17	20	15	15	30
1300	5:00	30	15	10	12	20	17	17	30
1400	6:00	20	12	10	10	20	17	17	30
1500	7:00	20	12	10	10	15	17	15	30
1600	8:00	20	12	10	10	12	17	15	30
1700	9:00	30	12	10	10	10	15	17*	30
1800	10:00	30	12	10	10	10	12	17	30
1900	11:00	30	15	10	10	10	10	12	17
2000	12:00	30	15	10	10	10	10	10	12
2100	1:00	30	17	10	10	10	10	10	12
2200	2:00	30	20	10	10	10	10	10	12
2300	3:00	17	20	12	10	10	10	10	10

MST	N	NE	E	SE	S	SW	W	NW	CST
	ASIA FAR EAST	EUROPE	S. AFRICA	S. AMERICA	ANTARCTICA	NEW ZEALAND	OCEANIA AUSTRALIA	JAPAN	
5:00	17	30	12	10	10	10	10	10	6:00
6:00	20	30	15	10	10	10	10	12	7:00
7:00	20	30	15	10	10	10	10*	15	8:00
8:00	20	30	15	10	10	10	10	17*	9:00
9:00	30	30	17	12	10	10	10	20	10:00
10:00	30	30	17*	12	12	10	10	20	11:00
11:00	30	30	17	15	15	12	12	20	12:00
12:00	30	30	17	15	17	12	12	30	1:00
1:00	30	30	17	15	17	15	15	30	2:00
2:00	30	30	17	17	20	15	15	30	3:00
3:00	30	17	17	17	20	15	17	30	4:00
4:00	20	15	12	17	20	17	17	30	5:00
5:00	20	12	10	12	20	17	17	30	6:00
6:00	15	12	10	10	20	17	15	20	7:00
7:00	15	12	10	10	17	17	12	30	8:00
8:00	12	12	10	10	15	17	15	30	9:00
9:00	15	12	10	10	12	15	17	30	10:00
10:00	17	12	10	10	10	12	17	30	11:00
11:00	20	12	10	10	10	12	12	30	12:00
12:00	20	15	10	10	10	10	10	17	1:00
1:00	20	15	10	10	10	10	10	15	2:00
2:00	30	17	10	10	10	10	10	12	3:00
3:00	30	20	10	10	10	10	10	12	4:00
4:00	30	30	12	10	10	10	10	12	5:00

EST	N	NE	E	SE	S	SW	W	NW
	ASIA FAR EAST	EUROPE	S. AFRICA	CARIBBEAN S. AMERICA	ANTARCTICA	NEW ZEALAND	OCEANIA AUSTRALIA	JAPAN
7:00	17	30	12	10	10	10	10	15
8:00	20	30	15	10	10	10	10	17*
9:00	20	30	15	10	10	10	10	20
10:00	20	30	15	10	10	10	10	20
11:00	30	30	17	12	12	12*	12	20
12:00	30	30	17	12	15	12	12	30
1:00	30	30	17*	15	17	15	15	30
2:00	30	30	17	15	20	15	15	30
3:00	30	30	17	15	20	15	15	30
4:00	30	17	15	17	20	17	17	30
5:00	15	15	10	17	20	17	17	30
6:00	15	12	10	12	20	17	17	20
7:00	12	12	10	10	20	15	15	20
8:00	15	12	10	10	17	15	12	30
9:00	17	12	10	10	12	15	12	30
10:00	20	12	10	10	12*	17	15	30
11:00	20	12	10	10	10	17	15	30
12:00	20	12	10	10	10	15	17	30
1:00	30	12	10	10	10	12	17	30
2:00	30	15	10	10	10	10	12	17
3:00	30	15	10	10	10	10	10	15
4:00	30	17	10	10*	10	10	10	12
5:00	30	20	10	10	10	10	10	12
6:00	30	30	12	10	10	10	10	12

days of very high MUF (solar flux) conditions. Usable distances on these bands should be somewhat greater than that achieved on 80 at night.

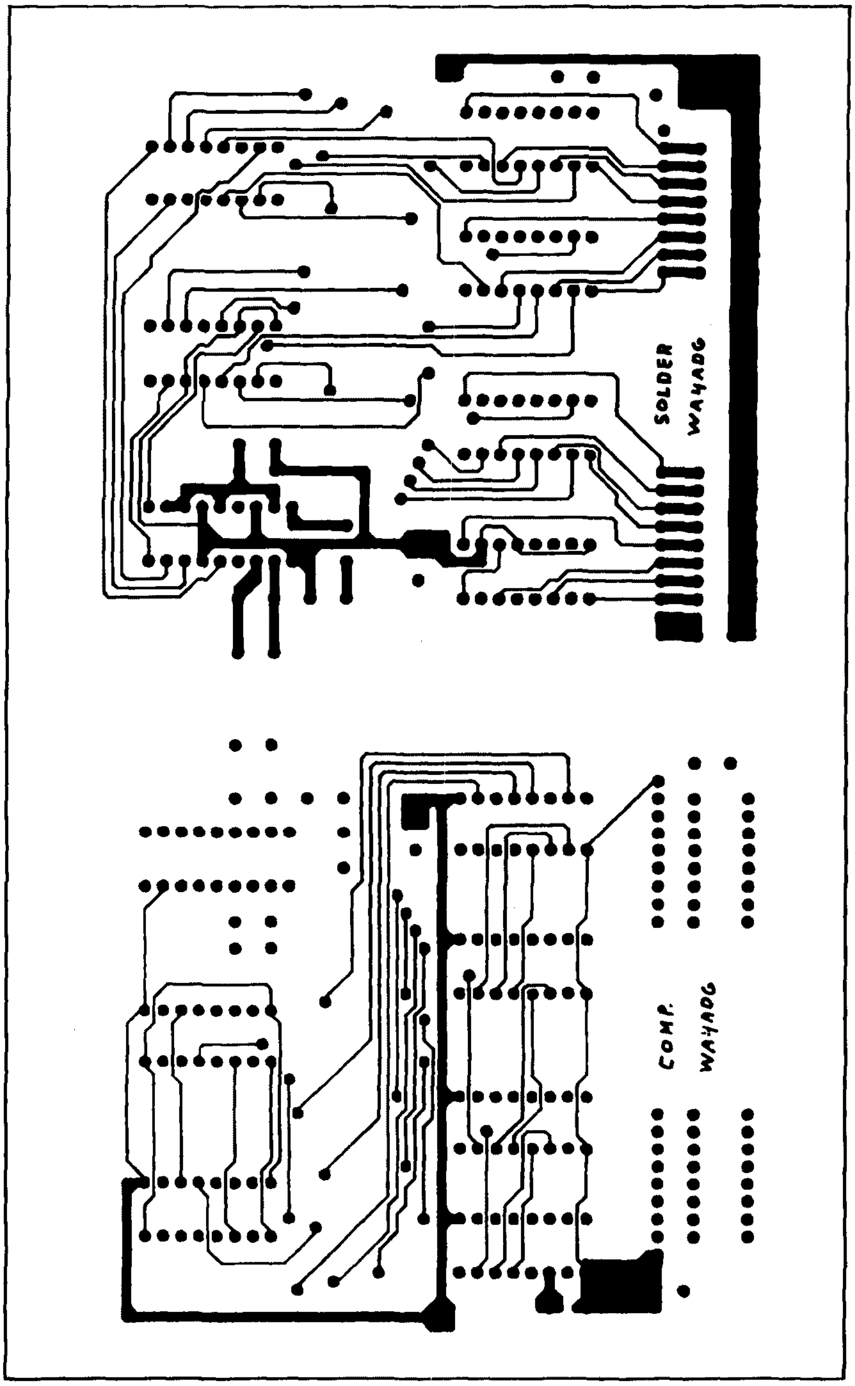
Forty, 80, and 160 meters, the nighttime DXer's bands, will open just before sunset and last until sunrise on the path of interest. Except for daytime short skip signal strengths, high solar flux values have little effect. Geomagnetic disturbances, more evident during the equinoctial periods, cause signal attentuation and fading on polar paths. Noise increases noticeably on these lower frequency bands in the coming months.

*Using two bands at once ensures a good signal during disturbances or interference.

Short Circuit: WA4ADG, December 1989

The pc board foil and component sides layout shown in Figure 1 on page 31 are incorrect. Here is the corrected artwork.

Please note that the pc board is double sided with all components mounted on the side designated COMP. The author did not use a pc board with plated through holes, but rather soldered all components on both sides where applicable. Similar pc boards can be obtained from FAR Circuits, 18N640 Field Court, Dundee, Illinois 60118 for $7.70 each plus $1.50 shipping and handling.



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DEADLINE 15th of second preceding month.

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BEGINNER'S RADIO CLEARINGHOUSE. On a space available basis, we are going to offer you, OUR SUBSCRIBER, free of charge, a chance to find a home for your used equipment with a new Ham. Please send us a short description of what you want to sell along with price, name, address and phone number. We'll run it once in a special section of the classified ads under the heading of BEGINNER'S RADIO CLEARINGHOUSE. Please limit your ad to 20 words or less.

FOR SALE: Heathkit Audio Generator, Hoathkit RF Generator, Heathkit Condenser Checker, RCA 3" oscilloscope. $200. WA2MCF, Charles Makish, 137-06 Laburnum Ave, Flushing, NY 11355. (718) 445-0558.

WANTED: Ham equipment and other property. The Radio Club of Junior High School 22 NYC, Inc. is a nonprofit organization, granted 501(C) (3) status by the IRS, incorporated with the goal of using the theme of Ham Radio to further and enhance the education of young people nationwide. Your property donation or financial support would be greatly appreciated and acknowledged with a receipt for your tax deductible contribution. It's easier, faster, and usually more profitable to donate than to sell. Most important, you're helping. Join us on the "Classroom Net" 7.238 at 1200 UTC and 21.395, 1330 UTC daily. Write us at: PO Box 1052, New York, NY 10002. Round the clock hotline: (516) 674-4072

TEST EQUIPMENT—Reconditioned (guaranteed), terrific prices, NBS calibration available. Visa, M/C, Amex, COD. Electronic Surplus Inc, 4350 Town Plaza, Suite 203, Houston, TX 77045. (713) 7;28j-9718.

FREE LIST of low cost ham equipment. Write to Jim Braddy, WA4DSO, 3037 Audrey Drive, Gastonia, NC 28054.

WANTED: BUY & SELL. All types of Electron Tubes. Call toll free 1-800-421-9397 or 1-612-429-9397. C & N Electronics, Harold Bramstedt, 6104 Egg Lake Road, Hugo, MN 55038.

RADIO SHACK Color Computer Ham Software and Hardware. Free catalog. Dynamic Electronics, POB 896, Hartselle, AL 35640. (205) 773-2758.

HAM SOFTWARE IBM/Compatibles 10 disks $26.95. MC/VISA/Discover. N5ABV EAPCO/H, Bx 14, Keller, TX 76248-0014. (817) 498-4242. 1-800-869-7208.

PC DRILL BITS sizes 55, 56, 58, 60, 62, 65, 66, 68, 70. $1.25 each or 5 for $5. Free shipping. S Systems, Box 1724, Tempe, AZ 85280-1724.

KENWOOD TS-930-S/AT transceiver $995. 55 foot Tristao tubular tower, completely refurbished, extended height 62 feet, $595. Tri-Ex MC-50 toer motor control kit $345. Tie-Ex TA54-L tilt over accessory $345. Custom built 6 meter half kilowatt, lo drive $395. Collins 75S-3 with AM, CW, SSB filters $375. Custom built 1296 MHz transvertor 80 watts out, .5 DB front end, 28 MHz IF, $595. Rich Shideler, 139 W. Walnut, Visalia, CA 93277. (209) 564-3960.

VIBROPLEX COLLECTOR'S GUIDE. W1IMQ's new illustrated reference includes bug identification, dating tips, complete patents, more, 87 pages, softbound. *Only $14.95 plus* $2.00 shipping U.S. (DX, $5.00 shipping). Mass. add 75 cents tax. Artifax Books, Box 88-HR, Maynard, MA 01754.

10 Mtr FM Hy-Gain CB Board with instructions $9.95. FM DET Kit $7.00. 40 channel switch $5.00. Crystal $5.00. SQ/VOL pots $2.00/pair. All tor $24.95 plus $2.00 s/h. MORNING DISTRIBUTORS, PO Box 717, Hialeah, FL 33011. (305) 884-8686.

GREAT FAMILY HOME on hill in Worcester, MA. Large private lot, 4 bedroom colonial, 2 baths, 2 enclosed porches, eat-in kitchen, formal dining, living room, cozy den, fireplace, double garage, built-in ham shack. $152,000. Call (617) 894-3943 evenings.

CHASSIS AND CABINET KITS. SASE. K3IWK. 5220 Harmony Grove Rd, Dover, PA 17315.

RTTY JOURNAL published 10 times per year for those interested in digital communications. Read about RTTY, AMTOR, MSO'S, PACKET, RTTY DX and Contesting. Plus technical articles concerning the digital modes. $12.50 per year (foreign higher). RTTY JOURNAL, 9085 La Casita Ave, Fountain Valley, CA 92708.

CUSTOM MADE EMBROIDERED PATCHES. Any size, shape, colors. Five patch minimum. Free sample, prices and ordering information. HEIN SPECIALTIES, Inc., 7960 SW Manitou Trail, Glen Arbor, MI 49636. (616) 334-4385.

IBM-PC RTTY/CW. New CompRtty II is the complete RTTY/CW program for IBM-PC's and compatibles. Now with larger buffers, better support for packet units, pictures, much more. Virtually any speed ASCII, BAUDOT, CW. Text entry via built-in screen editor! Adjustable split screen display. Instand mode/speed change. Hardcopy, diskcopy, break-in buffer, select calling, text file transfer, customizable full screen logging, 24 programmable 1000 character messages. Ideal for MARS and traffic handling. Requires 256k PC or AT compatible, serial port, RS-232C TU. $65. Send call letters (including MARS) with order. David A. Rice, KC2HO, 144 N. Putt Corners Rd, New Paltz, NY 12561.

LET THE GOVERNMENT FINANCE your small business. Grants/loans to $500,000. Free recorded message: 707-448-0270. (KH5).

QSL SALE! 100 QSL cards $8. 200/$11, 1000/$33. Guaranteed correct! Shipping included. Write or phone for samples. Shell Printing, KD9KW, Box 50B, Rockton, IL 61072. (815) 629-2193.

CLOUGH-BRENGLE SIGNAL GENERATOR. Model 299A. Can someone please help me with schematic or full manual. W2BLL, RD2, Box 72, Boonton, NJ 07005.

BEAT THE COLD! Melbourne, FL QTH: 4:2:2 (3100 sq/ft) house, 1.6+ acres, 220V wired shack with coax races built in, RHON/25 pad, workshop, pool, sprinkler system, in the country no restrictions, many tall pines and oaks, easy commute to Cape Canaveral, close to fishing/beaches/shopping, SAE for details and photos, $209,000. N0DH/4.

AVANTEK ATF10135, $12.00, MMIC's, P.C. board, SASE: WA3IAC, 7148 Montague St, Philadelphia, PA 19135.

"HAMLOG" COMPUTER PROGRAM. Full features, 18 modules. Autologs. 7-band WAS/DXCC. Apple, IBM, CP/M, KAYPRO, Tandy, C128 $24.95. HR-KA1AWH, POB 2015, Peabody, MA 01960.

220 MHz REPEATER FOR SALE. Spectrum Communications SCR-1000A commercial grade repeater tuned lor 222.72 rcv/224.32 xmit, 35 watts $1,500. Wacom Products WP-652 duplexer tor above— $500. IC-03AT $225. IC-38A $289. 220 MHz Ringo Ranger $40. 220 MHz Cushcraft Trunk Mount $39. 220 MHz Cushcraft Mag Mount $39. 220 MHz Cushcraft 4-pole antenna $75. All equipment in perfect operating condition. Contact K5ZCO at (214) 680-0475 days or (214) 363-5858 evenings.

BATTERY PACK REBUILDING: Don't pitch it—mail it—for FAST—PROFESSIONAL service! Satisfaction guaranteed! ICOM: BP2/BP3 $19.95, BPS $25.95, BP7/BP8/BP70 $32.95. KENWOOD: PB21 $15.95, PB21H $21.95, PB25/25H/26 $24.95. YAESU: FNB9 $19.95, FNB10 $23..95, FNB4/4A $36.95, TEN-TEC: $24.95 "U-DO-IT INSERTS" ICOM: BP3 $16.95, BP5 $22.95, BP7/8 $27.50, KENWOOD: PB21 $12.95, PB24/25/26 $19.95, TEMPO: S1-15/SERIES $22.95, YAESU: FNB4/4A $32.95, FNB10 $18.95, AZDEN: 300 $19.95, "NEW PACKS" ICOM: BP5 $43.95, BP8S $58.95, YAESU: FNB2 $19.95, FNB10S/FNB12 $44.95, SANTEC: 142/1200 $22.95. TELEPHONE/PAGER/COMPUTER/COMMERCIAL PACKS—Free catalog. Add $3.00 shipping/order. PA +6%. VISA/MC add $2.00. CUNARD, RD 6, Box 104, Bedford, PA 15522. (814) 623-7000.

DIGITAL AUTOMATIC DISPLAY. Any Radio. Be specific. Large 45-cent SASE. GRAND SYSTEMS, POB 2171, Blaine, WA 98230.

450 MHz SPECTRUM ANALYZER. Adapted from Nov 85 QST article by Al Helfrick, K2BLA. Features: Three digit LED center frequency digital readout. 12 position calibrated Scan Width, 1 kHz to 50 MHz, switchable bandwidth; Wide 5 300 kHz, Narrow 5 10 kHz. *Use your low frequency scope for the* display portion. Switchable 10 kHz video filter. Variable IF attenuator, LOG output calibrated in 10 db steps. For complete kit, order #450-KIT $459.95 plus $4.50 s/h. For assembled and tested unit, order 450-AST, $799.95 plus $6.50 s/h. Calif residents add 6% sales tax. Foreign orders add 15% for shipping. A & A Engineering, 2521 W. LaPalma, #K, Anaheim, CA 92801. (714) 952-2114.

IMRA International Mission Radio Association helps missionaries. Equipment loaned. Weekday net, 14.280 MHz, 1-3 PM Eastern. Nine hundred Amateurs in 40 countries. Rev. Thomas Sable, S.J., University of Scranton, Scranton, PA 18510.

INTERESTED IN PUBLIC SERVICE? Join your Local Radio Emergency Associated Communications Team. In Pennsylvania call (717) 938-6943 or write REACT, 1160 Old Trail Rd, Etters, PA 17319.

RUBBER STAMPS: 3 lines $5.00 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

ELECTRON TUBES: Receiving, transmitting, microwave... all types available. Large stock. Next day delivery, most cases. DAILY ELECTRONICS, PO Box 5029, Compton, CA 90224. (213) 774-1255.

RECONDITIONED TEST EQUIPMENT $1.25 for catalog. Walter, 2697 Nickel, San Pablo, CA 94806.

COMING EVENTS

Activities — "Places to go . . ."

SPECIAL REQUEST TO ALL AMATEUR RADIO PUBLICITY COORDINATORS: PLEASE INDICATE IN YOUR ANNOUNCEMENTS WHETHER OR NOT YOUR HAMFEST LOCATION, CLASSES, EXAMS, MEETINGS, FLEA MARKETS, ETC, ARE WHEELCHAIR ACCESSIBLE. THIS INFORMATION WOULD BE GREATLY APPRECIATED BY OUR BROTHER/SISTER HAMS WITH LIMITED PHYSICAL ABILITY.

KENTUCKY: March 3. The 14th annual Glasgow Swapfest sponsored by the Mammoth Cave ARC, Cave City Convention Center, Cave City. Starts 8 AM. Admission $4. Tables $3. each. Exams will be given. Talk in on 146.34/95. For information N4HCO, 1379 Whites Chapel Road, Glasgow, KY 42141.

MISSOURI: March 9. 30th annual Amateur Radio Auction sponsored by the Jefferson Barracks ARC, Concordia Turner's Hall, 6432 Gravois, south St. Louis City.
MASSACHUSETTS: March 10. The Barnstable Radio Club's annual Hamfest, Oak Ridge School, Sandwich. Open 10 AM to buyers. 8 AM for sellers. Advance tables $8. At the door $10. Amateur exams given. Contact Don, WA1AIC (508) 778-5673 or Henry (508) 255-2818.

NEW JERSEY: March 10. Springfest '90 sponsored by the Shore Points ARC, Holy Spirit High School, Rt 9, Absecon. Doors open 9 AM. Admission $3. Sellers spaces $5 each. Talk in on 146.385/.985 and 146.52. For information write SPARC, PO Box 142. Absecon, NJ 08201.

NEW YORK: March 11. WECAFEST '90 sponsored by the Westchester Emergency communications Association. New location—Mount Pleasant Armory, Valhalla. 8:30 AM to 3:30 PM. Admission $4. Talk in on WECA repeaters 147.66/.06, 222.8/224.4, 442.475/447.475. For information contact Bob Wilson, N2DVQ or Sarah Wilson, N2EYX, 2 Soundview Avenue, Apt AS, White Plains, NY 10606. (914) 997-8491.

ILLINOIS: March 11. The 30th annual Hamfest sponsored by the Sterling-Rock Falls ARS, Sterling High School Fieldhouse, 1608—4th Avenue, Sterling. Doors open 7:30 AM. Tickets $3/advance; $4/door. Tables $5 incl elec. Talk in on 146.25/146.85 W9MEP Repeater. For information, tables or tickets contact Sue Peters, Sterling Rock Falls ARA, PO box 521, Sterling, IL 61081 or call AC (815) 625-9262.

MICHIGAN: March 17. 29th annual Michigan Crossroads Hamfest, Marshall High School, Marshall. Sponsored by the Southern Michigan ARS and the Marshall HS Photo Electronics Club. 8 AM to 3 PM. Tickets $3/door; $2/advance, please SASE. For information write SMARS, PO Box 934, Battle Creek, MI 49016.

NEW HAMPSHIRE: Interstate Repeater Society's Hamfest, Lion's Club, Lions Avenue, Hudson. 8 AM to 4 PM. Admission $2. Tables $10. Talk in on 146.85, 146.55. Contact IRS, PO Box 693, Derry, NH 03038. Wheelchair accessible.

TEXAS: March 17-18. The Midland ARC's annual St. Patrick's Day Swapfest, Midland County Exhibit Building, East Highway 80, Midland. 10-5 Saturday and 8-2:30 Sunday. Preregistration $5. $6/door. Tables $6 each. VE tests. For information and reservations contact Midland ARC, PO Box 4401, Midland, TX 79704.

NORTH CAROLINA: March 17-18. Charlotte Hamfest and Computerfair, Charlotte Convention Center, 4th and College Streets, uptown Charlotte. Sat 9-5, Sun 9-2. Tickets $5/advance; $7/door. Children under 12 free. Tables $12 advance only. Tickets and tables good for both days. For information and reservations write Charlotte Hamfest, PO Box 221136, Charlotte, NC 28222-1136 or call (704) 536-7373.

FLORIDA: March 17-18. The 20th annual North Florida Ham/Swapfest sponsored by the Playground ARC, Shrine Fairgrounds, Ft. Walton Beach. Doors open 8 AM both days. Talk in on 146.19/79 repeater. For information contact the Playground ARC, PO Box 873, Ft. Walton Beach, FL 32549.

OHIO: March 18. The Toledo Mobile Radio Associations's Hamfest, Lucas County Recreation Center, Key Street, Maumee. 8 AM to 5 PM. Admission $3.50/advance; $4/door. Talk in on 147.27 rptr, 442.85 reptr. Contact Ron Morris, WB8ZIM, 28141 Glenwood Rd, Perrysburg, OH 43551. (419) 666-8063.

CONNECTICUT: March 18. The Insurance City Repeater Club's annual Flea Market, American School for the Deaf, West Hartford. 9 AM to 2 PM. Admission $2. Tables $15. Talke in on 146.28/88. Contact Chuck Motes, K2DFS, 22 Woodside Lane, Plainville, CT 06062.

KENTUCKY: March 24. 11th annual Hamfest sponsored by the Lincoln Trail ARC, Pritchard Community Center, Elizabethtown. Admission $4/advance; $5/door. For tickets, reservations, information contact Chuck Strain, AA4ZD, PO Box 342, Vine Grove, KY 40175. (502) 351-1715.

NEW JERSEY: March 24. Flea Market sponsored by the Chestnut Ridge Radio Club, Education Building, Saddle River Reformed Church, East Saddle River Road at Weiss Road, Upper Saddle River. Donation $1. Tables $10 first; $5 each additional. Tailgating $5. Contact Jack Meagher, W2EHD. (201) 768-8360.

CONNECTICUT: March 24. Annual Auction of the Radio Society of Norwich, Uncasville VFW, Uncasville. Setup 9 AM. Auction starts 10 AM. Admission is free. Wheelchair accessible. Bring equipment to be auctioned. Talk in on 146.73 repeater. For information call KA1BB at (203) 739-8016.

OHIO: March 25. The 12th annual Lake County Hamfest, Madison High School, Burns and Middle Ridge Roads, Madison. 8 AM to 3 PM. Admission $4/door. $3/advance. 6' tables $5. 8' tables Talk in on 147.21/81, 222.90/224.50. Contact LCARA Hamfest, 5777 Fenwood Ct, Mentor-on-Lake, OH 44060. (216) 257-2036.

ILLINOIS: March 25. Annual LAMARSFEST sponsored by the Libertyville and Mundelein ARS, Lake County Fairgrounds, Rt 120 & 45, Grayslake. Doors open 8 AM. General admission $3/advance; $4/door. Swap tables $7. For information SASE to LAMARS, PO Box 751, Libertyville, IL 60048 or call Bob Dick, NY9E (708) 362-9634 after 7 PM.

NEW JERSEY: March 25. HAMCOMP '90, the 18th annual flea market, sponsored by the Delaware Valley Radio Association, New Jersey National Guard 112th Field Artillery Armory, Eggerts Crossing Road, Lawrence Township, Trenton. Handi accessible. 8 AM to 2 PM. Admission $3/advance; $4/door. Talk in on 146.07-.67. For information and space reservations write HAMCOMP'90, c/o KB2ZY, RD 1, Box 259, Stockton, NJ 08559. Please SASE.

WEST VIRGINIA: March 31. Hamfest and Computer Show sponsored by the Kanawha ARC and the Tri-County Ham Radio Club. 9 AM to 3 PM. Charleston Civic Center, Charleston. Admission $5. VE exams. Talk in on 146.86/28 and 146.52. Contact Betty Palmer, WC8S, PO Box 8400, South Charleston, WV 25303. (304) 346-1348.

NEW JERSEY: March 31. Rain or Shine. The Cherryville Repeater Association's Flemington Hamfest, Hunterdon Central High Schokol Field House, Flemington. 8 AM. Admission $4/advance; $5/door. Chuldren under 12 and unlicensed spouses free. Table reservations and info from Marty Grozinski, NS2K, 6 Kirkbride Rd, Flemington, NJ 08822. (201) 788-4080, 5-11 PM EST.

ILLINOIS: March 31. Rockford Hamfest, 90.5533—11th Street, Rockford. 8 AM to 3 PM. Tickets $3/advance; $4/door. For information call Joe Roling, N9HEZ (815) 3919-6995 or SASE to PO Box 10003, Rockford, IL 61131.

MICHIGAN: April 1. 32nd annual ARRL Hamfest/Swap-n-Shop sponsored by the South Eastern Michigan ARA, Grosse Pointe North High School, 707 Vernier Rd, Grosse Pointe Woods. 8 AM to 2 PM. Tickets $2/advance; $4/door. Tables $8/advance; $10/door. For information SASE to SEMARA SWAP, PO Box 646, St. Clair Shores, MI 48080-0646. (313) 323-4099.

INDIANA: April 7. The Columbus ARC Hamfest, Bartholomew County 4-H Fairgrounds, State Rd 11, Columbus. 8 AM to 2 PM. Talk in on 146.79—600 Hz. For information Marion Winterberg, WD9HTN, 11941 W. Sawmill Road, Columbus, IN 47201. (812) 342-4670.

MINNESOTA: April 7. Rochester Area Electronic and Computer Show, John Adams Junior High School, 1525 NW 31st St, Rochester. Doors open 8:30 AM. Talk in on 146.22/82. For information RARC, 2824 NW 24th St, Rochester, MN 55901.

ONTARIO: April 7. The 9th annual Durham Region Amateur Radio and Computer Flea Market sponsored by the South Pickering ARC, VE3SPG and North Shore ARC, VE3NSR, Pickering High School, church St, Pickering Village, Ajax. 9 AM to 2 PM. Admission $4. For information VE2WZ (416) 839-3711; VE2MVD (416) 668-7590.

COLORADO: April 7. Swapfest sponsored by the Longmont ARC, Boulder County Fairgrounds, Longmont. 8 AM to 3 PM. Admission $3. Tables $7. For information or reservations contact Bob Dornan, WA2EKU, 1106 Fordham St, Longmont, CO 80501. (303) 651-3613.

MASSACHUSETTS: April 8. The Framingham ARA is sponsoring a Flea Market and license exams for all classes. Framingham Civic League Building, 214 Concord St. Rt 126, downtown Framingham. General admission, 10 AM, $2. Early Bird admission, 9 AM, $5. Talk in on 147.15. For table information Jon, K1VVC, (508) 877-7166. Exam information Dick, WA1KUG, (508) 877-0568.

OPERATING EVENTS

"Things to do . . ."

March 17-18. The Piscataway ARC (PARC) will operate their annual special event commemorating the Voice of America Relay station, WBOU, which operated during WW II in the Bound Brook section of Piscataway, NJ. Members will use their own callsigns signing /VOA. CW- Novice portions. Phone- Lower third of General on 75, 40, 20, 15 M and Novice portion of 10M band. For certificate send #10 or 9x12 SASE with QSL to PARC, att: KB2UV, PO Box 1233, Piscataway, NJ 08854.

THROUGHOUT 1990 the Major Armstrong Memorial Amateur Radio Club (MAMARC) will sponsor events commemorating Major Edwin Howard Armstrong's achievements in the field of radio broadcasting. The club is seeking other Amateur operators around the world who are willing to research Major Armstrong's accomplishments and become official MAMARC special events stations. Major Armstrong was a pioneer responsible for the creation of Wideband FM and the inventor of the superhetrodyne receiver. If you are interested in participating and becoming an official MAMARC special event station contact Barry Groupp, N2HDW, MAMARC, c/o 100th Birthday Committee, PO Box 581, Alpine, NJ 07620. Please SASE.

March 2: Grand Island, Nebraska, ARC will operate ARS W0CUO to celebrate the annual return of the Sand Hills crane to the Platte River refuge. 0000Z March 2 to 2400Z March 4. SSB, CW, PKT, AMTOR and RTTY in all lower portions of General and Novice bands. For Certificate send QSL and SASE to ARC W0CUO, PO Box 642, Grand Island, NE 68802.

YOUTH LINK NET. Open to all Hams under age 18. Saturdays at 2000 UTC, 28.425 MHz. For more information contact Net Control, George Manning, WB5NMH, 602 Glendale St, Burkburnett, TX 76354.

FREE 1989-90 Florida *two meter repeater directories are currently being distributed by the Hernando County Amateur Radio Assn. of Brooksville, FL. Ask for one at any official Florida Welcome Center or SASE to Repeater Directory, Hernando County ARA, POB 1721, Brooksville, FL 34605-1721.*

AMATEUR EXAMS. March 10, May 19, July 14, September 8, November 17. St. Mary Medical Center, 3333 No. Seminary Street, Galesburg, IL 61401. 12 Noon to 2 PM. For information contact Larry Heller, KA9PCU, 1436 Brown Avenue, Galesburg, IL 61401. (309) 342-5977.

Monthly Ham Exams. The MIT UHF Repeater Association and the MIT Radio Society offer monthly ham exams, all classes Novice to extra: next-to-last Wednesday of each month, (March 21) 7:30 pm, MIT Room 1-150, 77 Mass Avenue, Cambridge, MA. reservation requested a couple of days in advance, walk-ins welcome; call the shack (617) 253-3776, or Nich Altenbernd (617) 437-0320. Exam fee $4.95. Bring copies of your current license (if any) and Certificates of Completion (if any), two forms of picture ID, and a completed form 610, available from the FCC: (617) 770-4023.

LAUREL ARC monthly (except December) Amateur exam sessions for all license classes. No fee is charged. Pre-registration is required. Call (301) 725-1212, Maryland Radio Center, 8576 Laureldale Drive, Laurel, MD 20707.

AMATEUR RADIO CLASSES: For those people interested in obtaining a Novice (basic level) Ham license or upgrading to Tech/General, the Chelsea Civil Defense, in cooperation with QRA Radio Club, will sponsor Amateur Radio Communications classes evenings at Chelsea High School starting MARCH 7, 1989. For more information write Frank Masucci, K1BPN, 136 Grove Street, Chelsea, MA 02150. Please enclose your telephone number.


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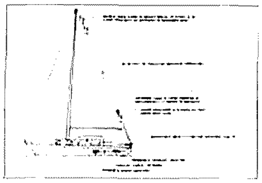
HAM RADIO

APRIL 1990
Volume 23, Number 4

Cover photo: Photo courtesy of the Hinckley Company, Southwest Harbor, Maine. Talaria, a Hinckley Sou'wester 51 sloop, is pictured here sailing off the Florida coast. Later renamed Skipjack, the sloop won the Bermuda One Two Race in 1989.



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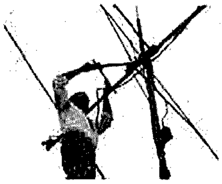

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FEATURES

DEPARTMENTS

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Backscatter

Life, Liberty, and the Right to Privacy

Have you ever wondered about a QSO in progress? Have you ever thought to yourself in the middle of a transmission, "Is this subject really appropriate for Amateur Radio?" Obviously, business communications are forbidden. But common sense also dictates that there are plenty of subjects best left to other methods of communication — like the telephone. Or are they?

An interesting case has come to light in Iowa concerning police monitoring of cordless telephone conversations. The November 29, 1989 edition of *The Wall Street Journal* reported that the Scott County Iowa Sheriff monitored a family's cordless telephone conversations over a 9-month period. This monitoring led to a 10-year sentence for conspiracy and theft. The family, in turn, has sued the sheriff for violating their constitutional right to privacy. In fact, they asked the Supreme Court to determine whether or not Fourth Amendment protection against unreasonable searches and seizures applied. Though the Supreme Court has decided not to hear the case, this story isn't over yet.

According to the *Journal,* lower court precedents don't favor Fourth Amendment protection for cordless telephones. They do apply, however, for more *technically sophisticated* cellular and wireless telephone conversations. Because the technology used for cordless telephones makes the calls easy to intercept, users shouldn't expect protection. The cellular telephone industry was able to ram legislation through Congress that could bring an all expenses paid "vacation" at the Allenwood, Pennsylvania Federal Correctional Facility for anyone hearing even the mere whisper of a cellular telephone conversation. To me this conjures up images of police raids on families huddled around TV sets scanning channels 70 to 83 in search of "neat" cellular conversations. While I shudder at the thought of people with paper pads and tape recorders ready to copy juicy tidbits of information, I hate to think that "Big Brother" might be looking over their shoulders while they do so.

This legislation is an example of a law which criminalizes an act that can't, in any way, be enforced. As a young ham, I feared the FCC. I was afraid that any indiscretion, even my beginner's sloppy fist, would bring the "radio police" to my door with a warrant for my arrest. In all practicality it wasn't going to happen, but I was still scared.

So what's the real scoop? Where should the line be drawn regarding what is and what isn't constitutionally protected in Amateur Radio conversations? For us radio Amateurs, the fight for privacy in communications will have little impact. However, it's always good to be reminded that when we talk, many are listening. A casual comment about going away for a weekend could result in a house break-in, or worse. There are millions of radios, scanners, and other types of equipment out there that can tune into Amateur communications.

It will be interesting to see if this matter is brought before the Supreme Court again. Hypothetically, if the case were to receive a favorable judgment, one could conjecture that all radio communications would enjoy protection. That would mean that you could call, but nobody could listen. Just think of it; the ham bands would be filled with endless CQs, and not one answer! *Is this scenario off the wall? Yes. Is it probable?* No. Well...maybe?

de NX1G

Comments

Lose some..win some!

Dear HR

I wanted to write to you and express my pleasure that you have picked up Bob Atkins, KA1GT, to write a new microwaves column. I was very sorry to see W1JR drop out, but I'm sure Bob's column will be every bit as interesting. This was the first column I would turn to in *QST* to read. So their loss is your gain!

Keep up the good work at publishing the best ham magazine available.

I might also suggest that possibly an ATV column would be nice. Perhaps W6ORG, Tom O'Hara, or KB9FO, Henry Ruh, would consider being contributing editors.

Again I'm very happy to see a microwaves column.

Dave Williams, WB0ZJP
St. Louis, Missouri

Early radar buff

Dear HR

"Understanding Over-the-Horizon Radar," *Ham Radio*, February 1990, brought back memories of field testing early warning radar SCR270 in the Signal Corps the summer of 1941 on Sandy Hook, New Jersey. It operated around 106 MHz and if the planes and thunderstorms were high enough it would reach out 100 miles or so. Not exactly over-the-horizon but a good radar application at the time. I understand from Dick Aspinwall, W7PV, that there is a museum where this equipment is on display along with the SCR584, an "S" band microwave radar Gunn director, that I later worked on.

Keep up the good work.

Wayne W. Cooper AG4R
Miami Shores, Florida

A good time to upgrade

Dear HR

I feel compelled to respond to N6SWA's "pet peeve" expressed in the February "Comments," regarding higher class licensees operating in the Novice/Technician 10-meter phone band.

I'm sorry, but I can't sympathize with him. Since the expansion of the 10-meter phone band, the majority of the "rarer" DX stations have been operating below 28.5 MHz. With the excellent propagation we are experiencing at the peak of the sunspot cycle, Novice/Techs have an unprecedented opportunity to take part in some of the best worldwide communications conditions we have had in many years. In six months it should be possible to work DXCC with 100 watts and a dipole with a little effort.

However, these conditions are not going to last, and will decline over the next couple of years. As an avid DXer, I am certainly not about to miss taking part in the peak activity and will operate wherever the DX is, with 100 watts if that will do, or a kilowatt if necessary.

Even under very good propagation conditions 28.3 to 28.5 has plenty of elbow room. And 200 kHz is more than most of us can operate on the other HF phone bands (except 15 meters). I have never heard QRM on 10 that comes close to what you'll find on 20 meters.

The best solution for the Novices and Techs that feel inhibited with their band allocations is, of course, to upgrade. I empathize with the Techs that have a problem with code (note that I don't have a 2X1 call) but with effort most people can make 13 wpm. Thanks to the question pool being "Bashed" a few years ago, the General Class (and higher) theory is now easily passed by anyone who is willing to read and memorize for a dozen or so evenings, with no real need for comprehension.

As further incentive, newcomers to the HF bands should be aware that it won't be very long until 10 meters will only be good for talking across town, and a sporadic E opening to another state will be a big deal. So "kwitcher-bellyakin," upgrade, and be ready to put up with some real QRM when 20 is the only band open and everyone is crowded into 200 kHz.

Bruce Sanborn, KB2WN
Rochester, New York

On the mark

Dear HR

I wish to comment on "Elmer's Notebook" in the January isue of *Ham Radio*. It is by far the best review on "resistance" I have ever read and has a very positive outline on the basics of the subject.

Thanks again for this and many other articles you have written. I look forward to "Elmer's Notebook" each month.

Stuart J. Tuma, W1QXS,
Melrose, Massachusetts

Dear HR

I receive *Ham Radio* magazine every month, and look forward to each magazine with anticipation. I am a project builder and enjoy all the articles.

My wife gives me the magazine every year as a Christmas present. You and your staff are to be congratulated on a very fine magazine. Keep up the fine work.

Phil Fuglsaug, WD0HXY
Imperial, Missouri

MARITIME MOBILES
NEED SIMPLE RIGS AND GOOD INSTALLATIONS

Sound tips keep maritime operations afloat

By Clyde Kirlin, WB6VPX, P.O. Box 2116, Walnut Creek, California 94595

Sailboats and Amateur Radio have been mates afloat for many years. Sailors have frequently explored ocean-going Amateur Radio as a hobby, but in recent years it has been used more as a lower cost substitute for a single sideband (SSB) marine radio. Today's technology makes it easier than ever for the multitasked sailor to use ham gear.

Dial twisting is a thing of the past. Memory bands now hold the frequencies we most commonly use. Microprocessor-controlled automatic antenna tuners quickly adjust for the chosen frequency. Push buttons, rather than analog rotation devices, simplify RF gain, mode selection, and tuning dial adjustments.

While operations are streamlined, good installation practices are often forgotten in the haste to "get on the air." Successfully fitting a radio to a sailboat involves creating a good ground plane, installing an exceptional antenna, and placing the equipment in proper relationship to antenna, ground plane, and power source. Additionally, the operator's transceiver selection process is often affected by brand preference, peer pressure, and impulse buying. Buyers frequently ignore the basic requirements and specifications that comprise a good receiver, as well as available power sources at sea.

Is the best really the best?

Take the case of my good friend Sam, who was about to fit his sailboat with an ICOM-781. Knowing he was a pragmatist at heart, I felt he had been swayed by the unsurpassed quality and features of the IC-781. Granted, it is a superb transceiver in a base station environment, but it wasn't designed for a sailboat. The IC-781's receiver performance is highly rated and I don't know of any radio that can equal the tricks performed with the IC-781's built-in keyer, filters, and twin passband tuning. Consider, however, what's required to maintain this performance level on a sailboat. Would Sam be using RTTY, AMTOR, and CW at sea? I doubt it. Good voice communications meet most objectives — no need for full break-in, noise blankers, and CW filters in that mode.

Because the 781 operates only on 120 volts AC, a 12-volts DC inverter or fuel-driven AC generator would be needed on a sailboat. A major maker of inverters agreed with my observations that a "demand-type" inverter wouldn't respond quickly enough to the fluctating power demands of a CW or SSB signal. And who wants to listen to a noisy generator on board while enjoying the pleasure of Amateur Radio? Further, the IC-781's automatic antenna tuner is designed for a 50-ohm output and not suited to match the varying impedance of an unloaded whip or random wire, as would be used on a sailboat.

Several other models using the same upbeat technology, but with simplified operation, stood at the head of the selection line. ICOM's compact IC-725 with its AH-3 automatic tuner* would be an excellent choice, as would economical performers from Kenwood† and Yaesu‡.

Many questions and some answers

Sam continued his quest for the best maritime setup with questions on how to create an antenna ground plane, the use of a mobile center-loaded whip or the boat's backstay as an antenna, and antenna length.

In my opinion, receiver specifications come first. The old adage applies, "If you can't hear 'em, you can't talk to 'em." Receiver sensitivities of less than 0.5 μV are quite common and most adequate for the quiet atmospheric environment of radio work at sea. Automatic antenna tuners will match almost any wire span that isn't near a half-wave in length in less than three seconds! All the operator needs to do is pick up the microphone and talk.

What makes a good installation?

A cruising sailboat is usually between 35 and 40 feet long, made of fiber glass, and rigged with one or two masts. Most sailboats carry a set of 12-volt DC batteries which are charged intermittently by the boat's engine alternator. Preplanning the overall installation layout is important. Because the antenna system will be similar to that of a quarter-wave ground plane vertical antenna, the antenna, antenna tuner, and ground plane must be in a reasonable vertical alignment. Mount the antenna tuner in a dry location in the lazarette area under the cockpit. Connect the antenna to the tuner here. The radiation current node

* ICOM America, 2380 116th Avenue, NE, Bellevue, Washington 98004.
† Kenwood U.S.A. Corporation, PO Box 22745, Long Beach, California 90801-5745.
‡ Yaesu USA, 17210 Edwards Road, Cerritos, California 90701.

TABLE 1

Antenna Length Calculations

Commonly used Amateur frequencies and the decimal portion of a wavelength that various antenna lengths represent

1. One-half wavelength = 492/Frequency (MHz)
2. One-half wavelength x 2 = one wavelength (W/L)
3. Antenna length/one wavelength (W/L) = decimal portion of W/L

	Antenna Lengths in Feet					
	23.00	30.00	35.00	40.00	45.00	50.00
Frequencies (MHz)	Wavelengths					
7.25	0.17	0.22	0.26	0.29	0.33	0.37
14.35	0.34	0.44	0.51	0.58	0.66	0.73
21.35	0.50	0.65	0.76	0.87	0.98	1.08
28.35	0.66	0.86	1.01	1.15	1.30	1.44

begins at the tuner output. Keep the portion of antenna wire inside the boat as short as possible to increase power radiation outside the hull.

Radiation resistance is composed of several factors, including ground resistance. Unfortunately, the antenna tuner usually is situated at some point above ground. Make the tuner think that it's sitting in sea water (ground) by using multiple copper ground straps between the tuner and the ground plane. Keep the straps as short as possible to reduce the ground resistance. The transceiver should be located at a convenient operating position, as close as possible to the batteries.

The antenna selection

Choose an antenna that fits the boat's rigging configuration, but is as long as possible. With today's high aspect ratio of tall mast and short boom, the boat's boom doesn't overhang the stern. This leaves an area for mounting a *23-foot vertical unloaded whip, like a Shakespeare* Model 390,* or a mobile center-loaded whip designed for a specific frequency. Of course, this leaves open for debate the question of insulating the backstay for longer antennas, creating higher efficiency on lower frequencies — versus the possibility of backstay failure in heavy weather.

Being safety minded, I opt for the 23-foot unloaded whip or the mobile center-loaded whip. With either of these, there are cautions to be considered. The 23-foot whip works well on 10, 20, and 40 meters. However, at 15 meters, it's close to one-half wavelength and won't load properly. The mobile whip must be mounted at the stern, using the metallic stern pulpit as a mounting base and possible ground plane. This creates a dangerous situation for the sailor attempting to change loading coils in a heavy seaway. Also, with the pulpit as a ground plane, a portion of the radiation — the lower 2 or 3 feet of the antenna mounting staff — is often wasted because it's sheltered by the ground plane itself.

How long is a vertical antenna?

The actual antenna length includes the distance from the tip of the antenna, the antenna lead-in to the tuner, through the ground lead from tuner to ground. Because antennas don't load very well at multiples of one-half wavelength, the formula in **Table 1** outlines the calculation of antenna lengths to be avoided.

For 10, 15, 20, and 40-meter operation, insulate the backstay to produce an antenna and ground lead length of 30, 40, or 45 feet. Note that a 35-foot antenna is close to a quarter wavelength at 40 meters — an ideal vertical antenna length. Yet the same antenna crowds the half-wavelength point on 20 meters. The total length from tip to ground *should not* be a multiple of one-half wavelength for any frequency being used.

The backstay

The backstay, when used as an antenna, must have strain insulators inserted in the wire. A good backstay insulator must be able to withstand 15-kV breakdown voltage and have the physical strength to endure the rigors at sea. Any backstay insulator work is best left to a professional rigging shop, for both selection and fitting. They must assume the responsibility for insulator failure and possible dismasting.

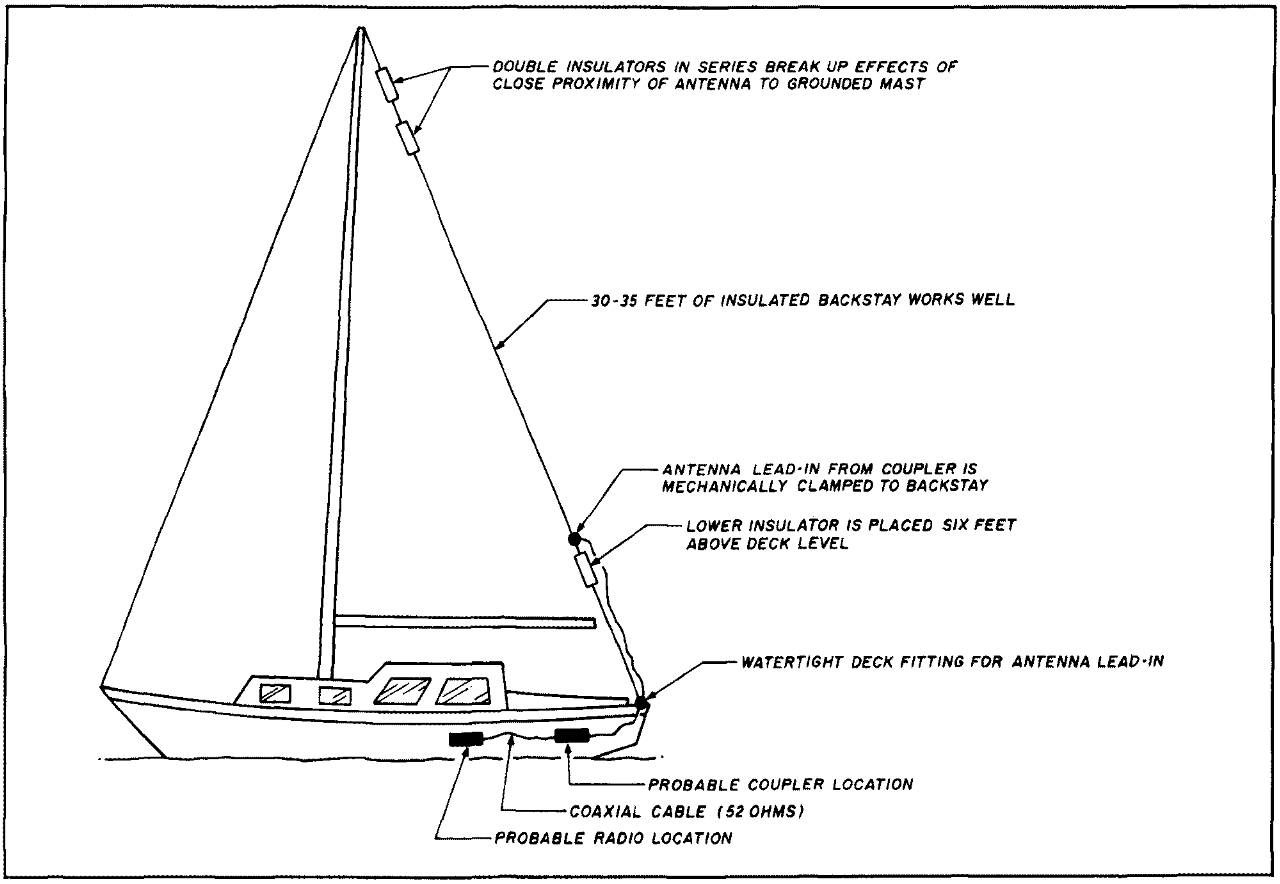
Have the lower insulator placed in the backstay at a height of 6 feet above the deck level, so those looking for a steadying grip at sea don't grab onto an antenna that's radiating power. Make sure there's high voltage insulation (25 to 30 kV) on the antenna lead from the coupler to the insulator.

The placement of the upper insulator depends on the distance between the point where the antenna tuner connects to the ground plane and the upper end of the desired antenna length. To offset any detrimental ground effects that a metal mast introduces to the antenna, a third insulator may be placed at the upper end of the backstay, within a foot or so of the masthead. "Living dangerously?" you ask. "A greater chance for dismasting," you observe. Yes, but this is why the insulator task is best left to professional riggers.

The antenna lead-in from the tuner to the backstay is a piece of high voltage insulated wire like GTO-15 neon sign wire or the polyethylene dielectric core of RG-8/U. One end of the lead-in is secured to the insulated portion of the backstay with U-bolts mounted close to the lower insulator. Wrap the U-bolt clamp with chafe protection. The tuner end of the lead-in is fed through a watertight deck fitting into the lazarette area near the antenna tuner (see **Figure 1**). *If you* have a metal deck, use a porcelain feedthrough insulator to keep the antenna lead-in as much as 6 inches from the

* The Shakespeare Company Antenna Group, PO Box 733, Newberry, South Carolina 29108.

FIGURE 1



Pre-planning the overall layout of the antenna system — the antenna, antenna tuner, and ground plane — for best results.

metal. Also, if the antenna lead-in is to be fixed in place on a metal hull, dress it 6 inches from all metal.

Creating the ground plane

What constitutes an adequate ground plane when afloat? Where should it be positioned? How large should it be? What material should be used? The only reasonable location for an antenna on a sailboat is far aft. The antenna, antenna tuner, and ground plane should be in reasonable vertical alignment. The only logical place for the ground plane is aft.

It's difficult to relate the ground plane area afloat with the number and length of radials used ashore. A coupling effect to the water exists if a ground plane is mounted low in the boat. The higher the ground plane is above the water, considering its necessarily limited size, the less efficient the system will be.

The best material for a ground plane on a boat is copper window screen, 16 mesh, found at most hardware stores. Cut the screen to the shape of the area in which it is to be mounted. Seam solder the wires of two edges at right angles to each other with a 2-inch wide, cold-rolled and annealed copper strap of 0.005-inch thickness. This way you'll know that all of the individual wires are a part of the ground plane. The copper strap readily allows other portions of the screen to be attached together with soldered joints. Secure the antenna tuner to the ground plane with several copper straps attached at different points on the ground plane. Look for copper straps at your local sheet metal supply house.

A good ground plane will lie high in the bilge (out of bilge water) in close proximity to the sea water. There's a degree of capacitive coupling to the sea water through the nonmetallic hull. As an alternative layout, use the underside of quarterberths — those berth areas generally found on either side of the cockpit or engine — for part of the total ground plane. Bond these portions of the screening with copper strap soldered to the bilge portion of the ground plane (see **Figure 2**).

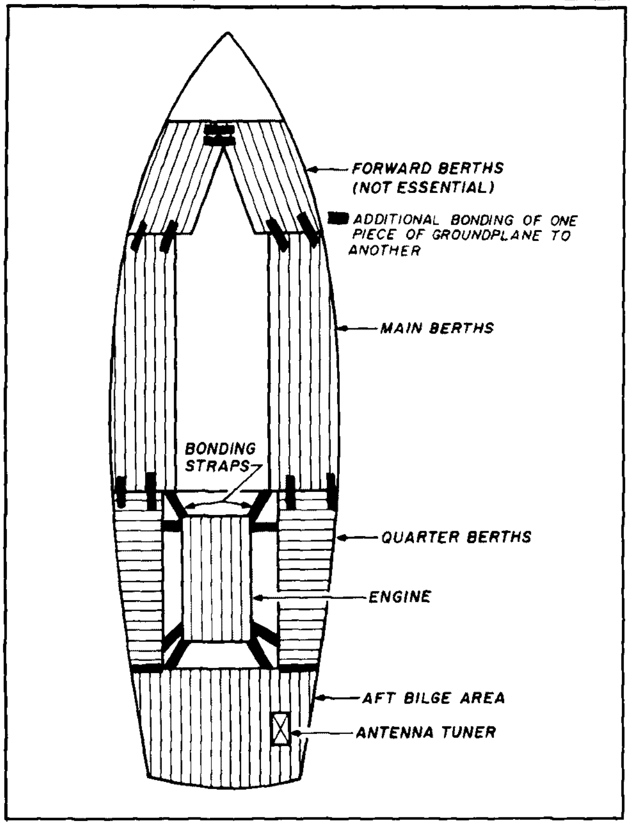
In **Figure 3** you'll see that tankage, metal hydraulic tubing, fire extinquisher piping, water lines, lifelines, toerails, engine, shaft, through-hull fittings, propeller, and rudder constitute the rest of the ground plane when bonded to other ground plane portions with copper strap and soldered joints.

With any boat installation, everything is a compromise. One could claim that 400 square feet of ground area is required, but a horizontal area that large is difficult to obtain on a boat. Yet a ground plane of 10 by 15 feet is the absolute minimum for good efficiency. The limitations of boat size and layout face you. Do the best you can.

Wrapping it up

With the antenna and the ground plane in place, and the antenna tuner positioned and attached to the ground plane,

FIGURE 2



What the antenna would see, if it were to search out possible candidates for ground plane or counterpoise areas. The areas of the cabin sole would add to this, but only if neatness could be maintained by covering screening with carpeting. However, in heavy weather cruising carpeting isn't always practical because it will get wet.

the transceiver can be wired to the boat's battery system. Good practice dictates that a dedicated electronic wiring bus be used to bring full power to an electronic distribution circuit breaker panel.

One hundred-watt transceivers draw 20 A when transmitting. The wire size between the battery and transceiver should be selected to limit the voltage drop to about 3 percent.

The American Boat and Yacht Council (ABYC)* recommends using AWG no. 6 wire if the length of the conductors from the battery to the transceiver **and back** to the battery is between 25 and 40 feet. Any longer wiring run will require AWG no. 4 wire.

Most manufacturers will provide a power cable less than 6 feet in length of an AWG no. 10 wire size. This may be used only if the battery is located within the length of cable provided. Otherwise, this cable should be shortened to less than 2 feet and terminated on a barrier terminal strip using crimp lugs. Use heavier wiring of a size that follows the ABYC dictates between the terminal strip and the distribution panel.

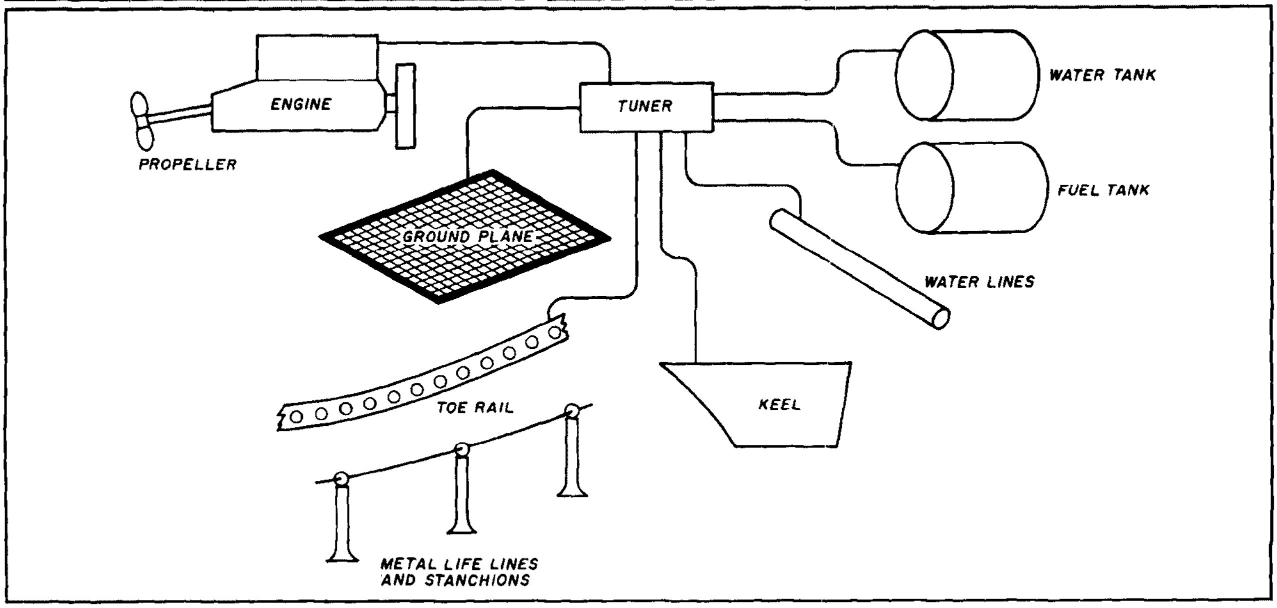
Wire the tuner and transceiver as directed and connect the two units with a piece of RG-58/U antenna lead-in cable to complete your system.

Time to get on the air!

All antenna and ground elements are secured in near vertical alignment. The ground plane is as large and as horizontal as you can make it, and located close to the bilge in the after section of your boat. Finally, due to your professional wiring, there's a full 12 volts DC at your rig when transmitting. If all the work has been done with care, there's not much left to do except to get on the air and listen to those gratifying comments from 10,000 miles away, "You're 20 over 9!" hr

* The American Boat and Yacht Council, PO Box 806, Amityville, New York 11701.

FIGURE 3



Make sure you bond to everything for a good ground plane, or counterpoise.

By Bob Atkins, KA1GT

OPTICAL RECEIVER

Last month I described the requirements for the transmitter end of an optical communciations link. This month I'll take a look at the other end of the link — the receiver.

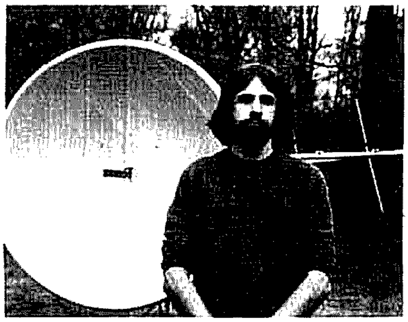
The optical receiver

Just as the equipment used in the laser transmitter has RF analogs, so do the components used in an optical receiver like the one shown in **Figure 1**. Probably the best analogy is that of a crystal set. The antenna of a crystal set would correspond to the lens (or mirror) of the optical receiver. The crystal set's tuned circuit corresponds to the optical filter, the crystal itself is analogous to the optical detector, and the headphones of the crystal set are replaced by the amplifier section of the optical receiver.

The first function of the optical receiver is to capture as much of the transmitted light as possible and to direct that light towards the optical detector. The function is the same as that of a telescope, with the receiver as the eye, so it's no surprise that an optical receiver closely resembles a telescope. Either a large lens or mirror can be used to capture the light. For visual use, the optics of a telescope system must be shaped with extreme accuracy. In fact, they must corresond with their designed shapes to better than one-quarter the wavelength of light (a few millionths of an inch). Though such optics would work well in an optical receiver, they are very expensive in large sizes. Rather than form a detailed image, the optics of the optical receiver need only concentrate the incoming light into an area less than or equal to the area of the optical detector; thus lower quality optics can be used to good effect.

The fresnel lens

One type of lens which has been used in optical receivers is known as the fresnel lens. This lens is comprised of concentric rings of prismatic elements, each of which bend the incoming light towards the lens's focal point as shown in **Figure 2**. The fresnel lens has advantages over conventional optics because it is much cheaper,

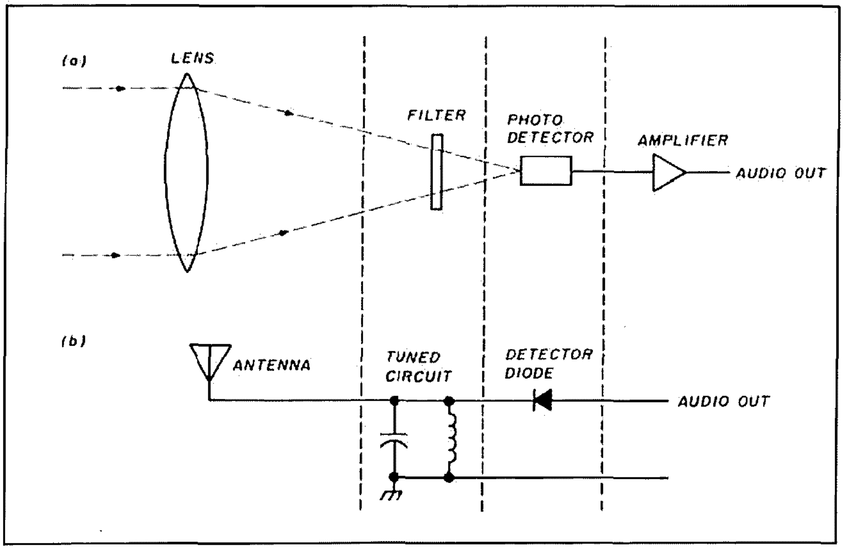


much thinner, and lighter. Unfortunately, the fresnel lens produces a very poor quality visual image. But, as I pointed out earlier, this may be of no consequence in an optical receiver. To compare prices, a 6-inch diameter optical lens might cost $700 and a 6-inch parabolic mirror might run $150, but a 12-inch fresnel lens (which captures 4 times as much light as the 6-inch optics) can be purchased for around $30.

Interference and absorption filters

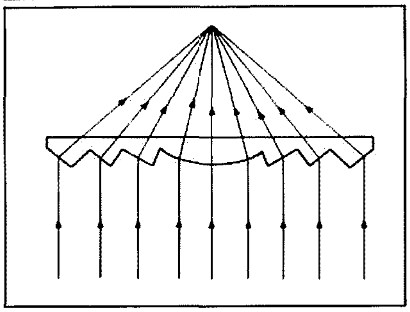
As mentioned last month, the light from a laser is monochromatic (i.e.,

FIGURE 1



(A)Schematic of a simple optical receiver. (B) Analogous components of a crystal set.

FIGURE 2



Cross-section of a fresnel lens.

composed of only one color). The background light entering the optical receiver will, in general, be composed of all visible colors plus some infrared and ultraviolet wavelengths. These unwanted background optical signals can be thought of as "optical QRM." A filter can be used to select out only the laser light, just as an RF filter is used to select one specific RF frequency.

There are two commonly used filter types — interference and absorption filters. Interference filters are composed of many layers of different dielectric

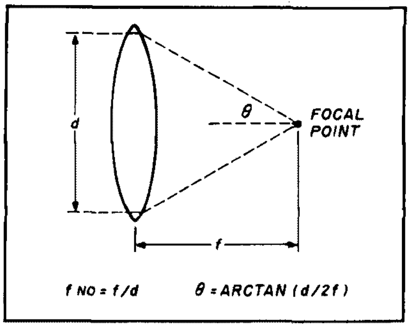
materials and work on the principle of the constructive and destructive interference of light. These filters have a very narrow passband, analogous to high Q cavity filters at RF. With interference filters, the wavelength of maximum transmission depends on the angle at which the light strikes the filter. This is of consequence when such a filter is used in conjunction with small f-number receiving optics.

The f-number of a lens is given by the focal length of the lens divided by its diameter. As you can see in **Figure 3**, light captured by the edge of such a lens will arrive at its focal point at a large angle (theta). This may be sufficient to cause the transmission maximum of an interference filter to shift to such an extent that light of the laser's wavelength is no longer transmitted. For a 10-nm wide filter at the He-Ne wavelength of 632.8 nm, light coming from the edge of an f2 lens will hit the filter at a 14-degree angle of incidence. This is enough to shift the filter peak response from 632.8 nm to around 628.4 nm, reducing the transmission by around 50 percent. Because only the light from the edges of the lens suffers this much attenuation, this configuration is acceptable. However, if an f1 lens is used (and this is not unusual for a fresnel-type lens), 632.8-nm light from the outer edges of the lens won't pass through the filter at all, and only light collected by the central region of the lens can be used.

What this all boils down to is that a 6-inch diameter f2 lens (12-inch focal length) will be as efficient at light collecting as a 12-inch f1 lens (12-inch focal length) when a 10-nm bandwidth interference filter is used. If you have a lens whose f-number is less than 2, there are two ways to use it efficiently. A small diverging (convex) lens can be placed before the filter. This converts the converging beam from the main lens to a parallel (or much less converging) beam, lessening the angle at which the light enters the filter. The other alternative is to use an absorption filter instead of an interference filter. Absorption filters absorb light of certain wavelengths and allow all others to pass through. In RF terms, they can operate like low pass, high pass, or bandpass filters of low to moderate Q.

In summary, the interference filter has the advantage of narrow bandpass (and hence high background light rejection) at the expense of sensitivity to incidence angle and cost. On the other hand, while absorption filters are low in cost and insensitive to incidence angle, they do not provide the background rejection of interference filters. For daytime operation the use of an interference filter is highly desirable, perhaps even essential, while nighttime operation is possible with an absorption filter, or even no filter if there's little or no background light.

FIGURE 3



Definition of the f number of a lens. Note: When the f number is small, angle θ is large. This can be a problem if an interference filter is used in the optical path (see text).

For He-Ne lasers, a 632-nm interference filter 0.5 inches in diameter with a 10-nm bandpass is available for $38 from Edmund Scientific (see last month's column). A red photographic filter (Kodak types 29, 26, or 25) would be acceptable as an absorption filter for use with He-Ne lasers. A 2-inch diameter filter should be available from local photographic equipment suppliers for less than $10.

After capturing and filtering the light, the next step is to convert the optical energy into electrical energy using an optical detector. There are two detector types of practical interest, the photomultiplier and the photodiode.

The photomultiplier is an electron tube device which contains a number of internal electrodes (dynodes). An overall supply voltage of around 1000 volts is split evenly between the electrodes. When a photon enters the tube it impacts on a photocathode and ejects an electron (this is known as the photoelectric effect). This electron is then accelerated to the nearest electrode by the potential difference between them. When it hits the electrode, it ejects several more electrons. These electrons are accelerated to the next electrode, where they strike it and emit even more electrons, and so on. Finally, after passage via the multiple electrodes, the initial electron release is effectively multiplied by up to 1 million times, and a measurable current is generated.

In contrast to the high voltage/amplification characteristics of the photomultiplier, the photodiode is a low voltage device with no intrinsic gain. It operates much like a solar cell, generating a small current when exposed to light. To be useful at low light levels, the current generated is converted to a voltage and amplified. This is accomplished most simply using an op amp configured as a transimpedance amplifier. Some photodiodes are available with a built-in high gain amplifier in the same package. This is a desirable configuration because, owing to the very high gains needed ($>10^5$), pickup of stray signals is a problem. These signals can be minimized by housing the photodiode and amplifier together in the same shielded package.

Choosing an optical detector

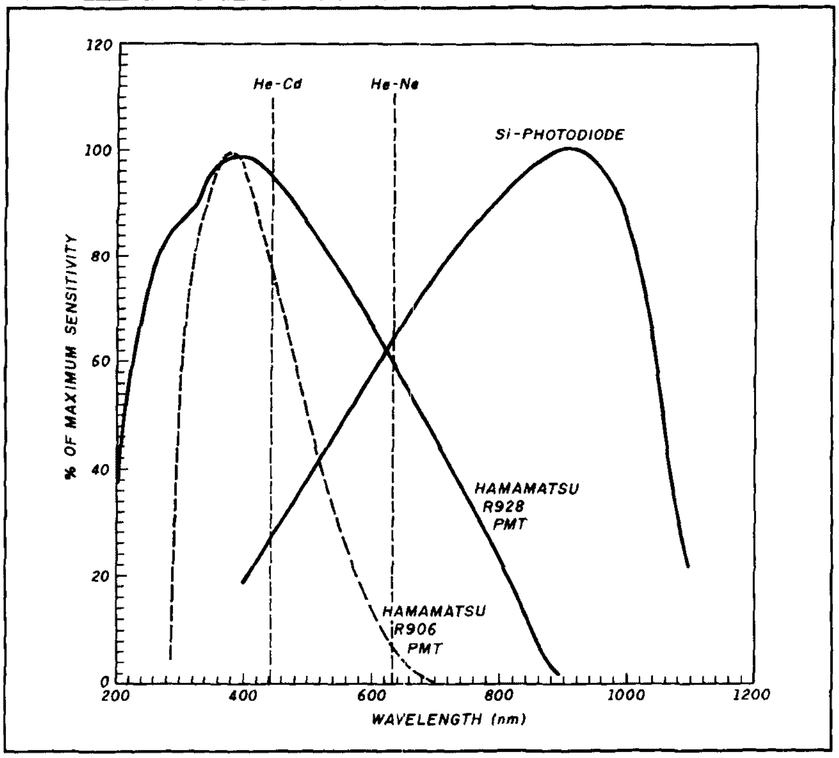
The choice of optical detector depends on a number of factors including spectral response, size, sensitivity, and convenience of use. Most photomultipliers are more sensitive in the blue region of the spectrum than the red. A typical tube might be 10 times more sensitive at 442 nm (He-Cd) than at 632 nm (He-Ne). Some tubes do have extended red response, but they are less common on the surplus market and more expensive. Photodiodes usually show maximum response in the near infrared at around 900 nm. They are usually about 1.5 times more sensitive at 632 nm than at 442 nm. The shapes of the response curves for several photodetectors are shown in **Figure 4**.

Detector area is also an important parameter. For maximum receiver efficiency, all the light collected by the lens must hit the active area of the detector. Using a fresnel lens will result in a larger image than would be obtained using high quality optics. It's evident, therefore, that a larger area optical detector is desirable because it will increase the probability that all the collected light will hit the detector. For the same reason, a large detector area also lowers the required pointing accuracy. A potential drawback of a very large area detector is that, given

reasonable imaging optics, it will collect light from a wider area than a small area detector (hence the easier pointing). This means that if there's a lot of background light, it could swamp out the laser signal. This shouldn't be a problem at night and/or when efficient narrowband filters are used. A typical photomultiplier has quite a large active area of around 1 cm × 2 cm.

Photodiodes are available in a range sizes from an active area of several square centimeters down to less than 1/10 of a square millimeter. Very small area detectors, like those of the Motorola MRD series (MRD 360 photodarlington, MRD 510 photodiode, and MRD 300 phototransistor) are quite sensitive and inexpensive ($2), and even include photodarlington transistor detectors which effectively show intrinsic gain. Unfortunately, for the reasons given previously, their extremely small active area (0.03" × 0.03" maximum) makes them a poor choice for most free space DX optical receivers. However, they are good and inexpensive devices for initial experimentation. Large area detectors are normally more expensive ($10 to $30), but will provide much better results in most cases. They do generate a little more internal noise, though in this application it shouldn't be enough to cause a problem.

FIGURE 4



Relative response curves for photodetectors.

Noise equivalent power

Ultimate detector sensitivity is limited by the intrinsic noise generated internally by the device. There is a measure of this noise known as the noise equivalent power (NEP). This is roughly comparable to the noise figure of a diode or transistor. NEP is defined as the amount of optical power required to generate a signal equal to the intrinsic noise level of the detector. Therefore, it's also the amount of optical power required to give a 0-dB signal-to-noise ratio. Because noise is a function of bandwidth, it's expressed in units of W/Hz$^{1/2}$. The NEP of a good photomultiplier is about 10 times lower than that of a good photodiode. Some typical NEP values for devices which would be suitable for use in Amateur optical receivers are:

Hamamatsu R906 photomultiplier
at 440 nm = 3e-16 W/Hz$^{1/2}$
(blue sensitive, $70)
at 632 nm = 6e-15 W/Hz$^{1/2}$

Hamamatsu R928 photomultiplier
at 440 nm = 1e-16 W/Hz$^{1/2}$
(red sensitive, $300)
at 632 nm = 2e-16 W/Hz$^{1/2}$

Hamamatsu S2386-8K photodiode
at 440 nm = 8e-15 W/Hz$^{1/2}$
(area = 33mm^2, $30)
at 632 nm = 3e-15 W/Hz$^{1/2}$

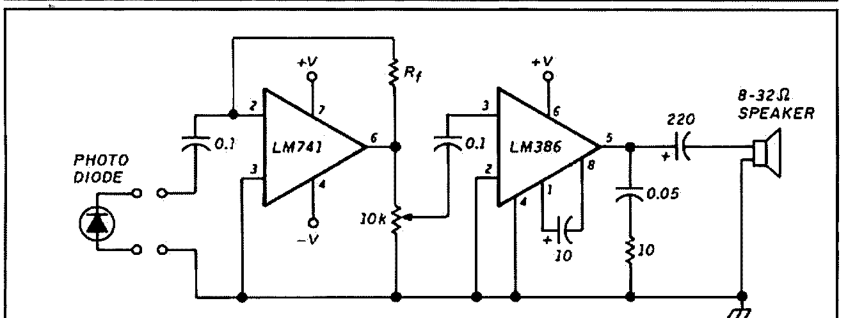
UDT-020D photodiode/amplifier
at 440 nm = 2.5e-14 W/Hz$^{1/2}$
(area = 20mm^2, $43)
at 632 nm = 5.1e-14 W/Hz$^{1/2}$

If the laser transmitter is tone modulated before keying (MCW or A2A modulation as described last month), all that's required to produce an audible CW signal is to amplify the output of the optical detector and feed the signal into an audio amplifier. A narrowband audio filter centered on the modulation frequency can be included for improved performance. If the laser is not tone modulated (CW or A1A modulation), and if DC coupling is then used throughout, the output of the photodetector after the transimpedance amplifier is a voltage level which rises and falls as the transmitter is keyed. To produce an audible signal, this voltage level must be used to key a tone on and off. This can be achieved by feeding the amplifier output into one input of a comparator. A reference voltage is then fed into the other comparator input and adjusted so the comparator output is low when the laser is off and high when the laser is on. The output of the comparator can then be used to key a tone on and off — or its digital state can be read directly by a computer.

Figure 5 shows the electrical schematic for a complete optical receiver suitable for the reception of keyed tone-modulated transmissions. If an integrated detector/amplifier is used, this would replace the first amplifier stage. Note that such detector/amplifiers are DC coupled, which means that their output saturates (reaches the supply voltage) with fairly low levels of incident light. For example, the EG&G HUV 2000B, using a ±15 volt supply, will saturate at about 3×10^{-7} watts of incident light energy when operating at maximum "gain." (Strictly speaking, "gain" is an inappropriate term because the amplifier is operating as a current-to-voltage converter.) In a well-constructed optical receiver,

and with a low level of background light, the detection of limit will probably be in the range of 10^{-10} to 10^{-12} watts of modulated laser energy — depending on the quality of construction, noise levels generated in the amplifier stages, the receiver bandwidth, and how well you can copy weak CW! For those interested in a detailed mathematical analysis of the noise components of optical detectors, check out the EG&G Photon Devices catalog cited at the end of this article.

FIGURE 5



NOTES

+V = +9V TO +12V

-V = -9V TO -12V

R_f = 100k - 1M THIS RESISTOR DETERMINES THE GAIN OF THE FIRST AMPLIFIER STAGE

A LOW NOISE OP-AMP MAY BE SUBSTITUTED FOR THE LM741 FOR IMPROVED PERFORMANCE

Schematic of the electronic stages of an optical receiver.

Suppliers

Here are a few sources for the components required to construct an optical receiver. The catalogs and data books referenced are free to commercial users, but there may be charges to individuals who order them. This is not an exhaustive list; inclusion or omission from the list does not imply endorsement or otherwise.

Hamamatsu — Manufacturer of photomultipliers and photodiodes. Large range of products includes PMTs from $70, photodiodes from $10. Minimum order $25. Will take credit cards. Catalogs available. Contact Hamamatsu, Order Service Department, 360 Foothill Road, Box 6910, Bridgewater, New Jersey 08807-0960.

United Detector Technology — Manufacturer of photodiodes. Good selection, from $10 for small photodiodes and $21 for small photodiode/amplifier combinations. Minimum order $100. Data sheets available. Contact UDT, 12525 Chadron Avenue, Hawthorn, California 90250.

EG&G Photon Devices — Manufacturer of photodiodes. Small photodiodes from $17, diode/amplifier combinations from $70. No minimum order. Catalog available. Contact EG&G Solid State, 35 Congress Street, P.O. Box 506, Salem, Massachusetts 01979-6526.

Motorola — Manufacturer of small area optoelectronic devices (photodiodes, phototransistors, photodarlingtons), mostly in the $2 to $15 price range. Optoelectronics Data book available from Motorola Literature Distribution, P.O. Box 20912, Phoenix, Arizona 85036. Devices available through distributors (e.g., Newark, Hamilton-Avnet).

See last month's column for suppliers of optics and filters.

Next month, in the final article on optical communication, I'll discuss the factors (principally scattering) which influence optical communication range and I'll show you how to calculate the DX potential of a laser communication system. Please send comments, questions and any microwave activity reports to Bob Atkins, KA1GT, 103 Division Avenue, Millington, New Jersey 07946.

A 220-MHZ RECEIVING ADAPTER

By Peter J. Bertini, K1ZJH, 20 Patsun Road, Somers, Connecticut 06071

Catch the wave — 1-1/4 meter waves, that is — with your 2-meter handheld and this simple converter! This weekend project allows casual 1-1/4 meter monitoring on your 2-meter handheld or VHF scanner. The self-contained converter is small enough to mount directly on the antenna jack of most BNC outfitted handhelds. Many converter designs have been published, but I wanted a circuit that didn't need an etched pc board. I also wanted to make this project as simple as possible!

Works both ways!

The converter uses a passive diode mixer and is bilateral; you can also receive 2-meter signals on a 220-MHz receiver! In this mode the converter is connected "backwards." My service monitor, an older Cushman CE-3, covers 144 MHz but not 220. With this converter, I can use the monitor to perform accurate frequency/deviation monitoring and limited signal generation on the 1-1/4 meter band (more on this later).

A VHF converter may use several stages. You might find an LO oscillator with one or two multiplier stages, a mixer, and usually an RF amplifier. This is hardly something you could build in a weekend or would want sitting on top of your 2-meter HT! I make do with only a mixer and oscillator. The mixer has fairly high losses and the ultimate sensitivity is determined by the 2-meter IF receiver.

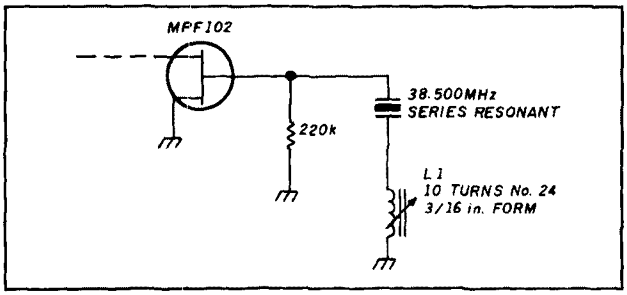
Construction starts with plastic case

The converter is housed in a plastic Radio Shack project case. A matching perfboard with 0.1-inch spaced component mounting holes comes with the case. A VHF converter would normally be built on an etched pc board. I avoided this by mounting a piece of sheet brass, slightly less than 0.7 inches wide, transversing the center length of the perfboard. All component leads going to ground connect directly to this strip. The brass stock is available at hobby shops and can be cut easily with tinsnips. A thickness of 0.01 inch will do. When mounting the trimmer capacitors, connect the rotor plates to ground. This lets you use metallic tuning tools during alignment, except when you're working with the trimmer on the oscillator secondary winding where both plates are at RF potential.

The BNC connector ground lugs are connected to the ground plane. This also holds the finished converter board in place. Remove the two plastic standoffs inside the case, so the converter board will sit flush with the case bottom. Carefully break away the standoff material with a small pair of needle-nose pliers. You may clean any remaining rough edges with a penknife. With care, everything — including the 9-volt battery — will fit inside the case. Parts layout isn't critical.

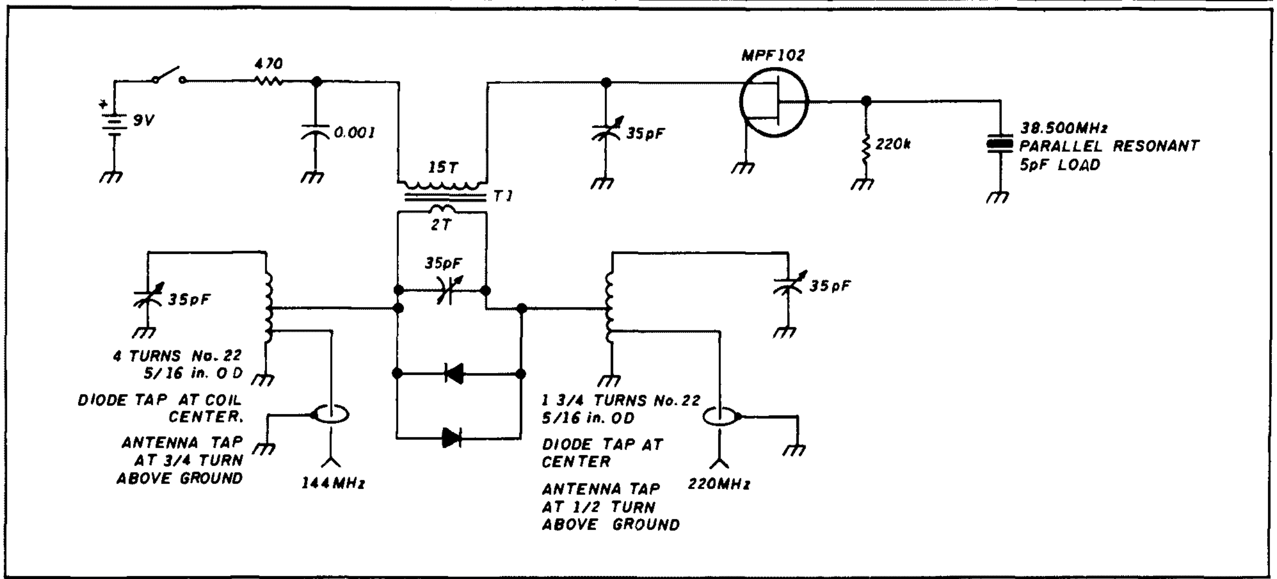
I used chassis mount BNC female connectors. You'll need a double male BNC adapter to connect the converter to most handhelds. You can also use a cable end male connector on the case for the radio connection. Use the rear cable nut to mount the connector through the enclosure. Prepare a short piece of bare hookup wire and solder it to the male pin of the BNC connector. Insert the pin so it's properly seated in the connector. Several drops of epoxy will keep the pin from pushing back when mated.

FIGURE 1



Variation of oscillator circuit uses series-resonant crystal with adjustable series inductance to permit precise setting of oscillator frequency.

FIGURE 2



Receiving converter adapter circuit. Oscillator is shown with parallel-resonant crystal cut for 5-pF load and is accurate to within a few hundred cycles. Coil details are given in the test and parts lists.

PARTS LIST

Capacitors

4 Trimmers, between 30-40 pF maximum, plastic or ceramic miniature.
1 Ceramic disc, 25 or greater working volts 1000 pF (0.001 μF)

Coils

1 220-MHz antenna coil, 1-3/4 turns no. 22 tinned wire, air wound. 5/16-inch OD, center tap for diode connection. Antenna tap is at 1/2 turn above ground.
1 144-MHz antenna coil, 4 turns no. 22 tinned wire, 5/16-inch OD, center tap for diode connection. Antenna tap is at 3/4 turn above ground.
1 L1 warping inductor. About 12 turns AWG no. 24 magnet wire solenoid wound on 1/4 to 3/16-inch slug-tuned form. Used only with series resonant cut crystal.

Crystal

1 38.5000-MHz, third overtone cut, HC-18u wire lead case.
Note: order ICM factory no. 4713601 for general use. Parallel resonant, 5-pF load.

Order ICM factory no. 471360 if it is desirable to set frequency. Series resonant mode. Used with series inductor to set frequency.

Diodes

2 High speed switching, silicon 1N914, 1N4148 or hot carrier type.

Project case

1 Radio shack 270-283 or similar

Resistors

1 1/4 watt carbon or film composition, 20 percent or better
1 470 ohm
1 220 k

Transistor

1 FET, MPF102 or 2N5486

Transformer (T1)

1 Toroid, T-37-6, yellow SF mix, 0.37-inch OD primary winding 15 turns AWG no. 24 enamel magnet wire,secondary winding 2 turns AWG no. 24 enamel magnet wire wound in phase over primary winding.

Oscillator is only active stage

The third overtone oscillator is the only active stage in this converter. Because I wanted 223 MHz to correspond to 146 MHz, I needed an LO frequency of 77 MHz. The third overtone oscillator operates at one-half the LO frequency. For casual monitoring, use an International Crystal catalog no. 4713601 38.5000-MHz crystal. It will be within several hundred cycles and will work fine. This parallel resonant crystal is cut for a 5-pF load.

For critical applications — like using the converter with a service monitor to set frequencies — I suggest an International Crystal catalog no. 471360 38.500-MHz crystal. This series resonant crystal can be set, or "warped," to exactly 38.5000 MHz by placing a small variable inductor in series with the crystal. See **Figure 1** for details. About ten turns of no. 24 wire on a 3/16 to 1/4-inch slug-tuned form will suffice. Be careful, too much inductance will cause spurious oscillations; some cut-and-try experimentation may be necessary.

Innovative mixer simplifies circuit

The mixer is driven directly by the 38.5-MHz oscillator. This unique passive mixer stage uses two diodes paralleled back to back (see **Figure 2**). Each diode conducts on alternating crests of the LO injection. This halves the required LO injection and eliminates the need for a frequency doubler. While this isn't the ultimate mixer, it serves my needs. I spent a lot of time optimizing the mixer circuitry, particularly when trying to couple the LO energy into the diodes. The best performance was achieved with the diodes connected across the output link over the oscillator tank circuit; however, losses were still very high. I discovered (entirely by accident!) that by resonating the coupling link to 2 meters the mixer losses dropped by several dB — hence the trimmer capacitor across the link winding. Why this improves performance I honestly don't know, but I surmise the coupling link loads the mixer diodes until they are transformed into

a high impedance parallel resonant circuit. It's possible that dividing the coupling link into two series parallel resonant circuits, one at 144 MHz and the other at 220 MHz, might further reduce the mixer losses.

I tried both hot carrier and high speed switching diodes, and noted no discernible difference. The HCDs may offer the advantage of better performance with lower LO drive as the battery voltage falls off. Diodes like the 1N914 or 1N4148 work well in this mixer.

Mixer limitations

The mixer has 12 dB of insertion loss. This is a little high and is a fair compromise between cost and circuit simplicity. Despite the losses, I achieved a respectable 0.5-μV sensitivity with the ICOM 2-AT. Repeaters within 20 miles of my QTH are full quieting with only a rubber duck antenna. With just one tuned stage used at each mixer port, unwanted converter image responses and high levels of LO and related harmonic energy leakage may occur. The converter offers about 20-dB attenuation port to port. Thus, for example, a strong signal on 146.82 MHz may compete with a repeater on 223.82 MHz.

Where to find the parts

I made every effort to use common parts. Many of the components can be found at the local Radio Shack. Most of the other parts (trimmer caps, toroids, FETs, etc.) can be obtained via mail order. I recommend you try Radiokit or Circuit Specialists. I used handwound coils and capacitive tuning, and avoided the high cost of slug-tuned inductors, which are hard to find anyway. I also used inexpensive 35-pF plastic trimmer capacitors. Radio Shack currently offers 50-pF trimmers. You can use these in a pinch, but you may find the tuning overly sharp. The airwound coils were made from no. 22 gauge hookup wire stripped of its insulation and wound over a pencil. The wire gauge isn't especially critical; neither are the coil dimensions — within reason.

Alignment

Oscillator tuning is performed by monitoring the current drawn by the converter. When oscillating, the current will drop sharply — typically from 8 mA to about 2 to 4 mA. The trick here is to trade maximum oscillator output for reliable oscillator startup when power is removed and reapplied. The oscillator should maintain oscillation and start reliably with voltages as low as 6 volts. If you use it, set the warping inductor to minimum inductance (slug out) during testing. You may use a frequency counter to set the oscillator to exactly 38.500 MHz, or you can simply adjust for minimum distortion on monitored signals. The oscillator's fourth harmonic can also be zero beat to 144.000 MHz on an SSB transceiver.

Tune the 144 and 220-MHz tuned stages for best sensitivity. You can use a signal generator or strong local 220-MHz signal. If a peak occurs at minimum or maximum capacity, try spreading or compressing the coil windings to compensate. Finally, you can tweak the oscillator tank slightly for best sensitivity. You may have to retweak the input circuit a bit if you change antennas. Use a 220-MHz antenna for best sensitivity.

I use a 9-volt alkaline battery for power. The battery will last a long time with only a 2 or 3-mA draw. For continuous monitoring use a wall plug-type power supply. The converter is forgiving of accidental transmitter keying, but it is not burnout proof. Always set the handheld to low power before connecting the converter.

Using the converter with a service monitor

Need 220-MHz coverage on your service monitor? This converter makes that possible. First, calibrate the converter. Set the preselector and CE-3 to monitor 144.000 MHz. Connect the 2-meter port to the preselector input and adjust the oscillator frequency to exacly 144.000 MHz. This sets the converter to the limits of the CE-3's own calibration.

To monitor a 220-MHz signal, connect the converter 2-meter port to the preselector input. Set the Cushman to the 2-meter frequency corresponding to the 220-MHz frequency to be monitored. You can now make very accurate 220-MHz frequency and deviation checks. Connecting the 2-meter port to the signal generator output will produce 220-MHz signals. Although the generator calibration won't be very accurate because the conversion losses will affect the readings, you can still perform meaningful relative comparisons and receiver peaking. hr

Crystals

International Crystal Manufacturing Company
P.O. Box 26330
701 West Sheridan
Oklahoma City, Oklahoma 73126-0330

Case, resistors, miscellaneous components

Radio Shack, local retail outlet

RF part, trimmers, caps, resistors, toroids, etc.

RADIOKIT
P.O. Box 973
Pelham, New Hampshire 03076

Jameco Electronics
1355 Shoreway Road
Belmont, California 94002

BCD Radio Parts Company
P.O. Box 119
Richardson, Texas 75080-0020

Circuit Specialists
Box 3047
Scottsdale, Arizona 85257

Toroids

Amidon Associates
12033 Otsego Street
North Hollywood, California 91607

A well-stocked junkbox is the best source of parts. I suggest writing the above companies for current catalogs; prices, minimum orders, and stock vary continually.

DIGITAL SIGNAL PROCESSING
PART 1: THE FUNDAMENTALS

By Bryan P. Bergeron, NU1N, 30 Gardner Road, Apt. 1G, Brookline, Massachusetts 02146

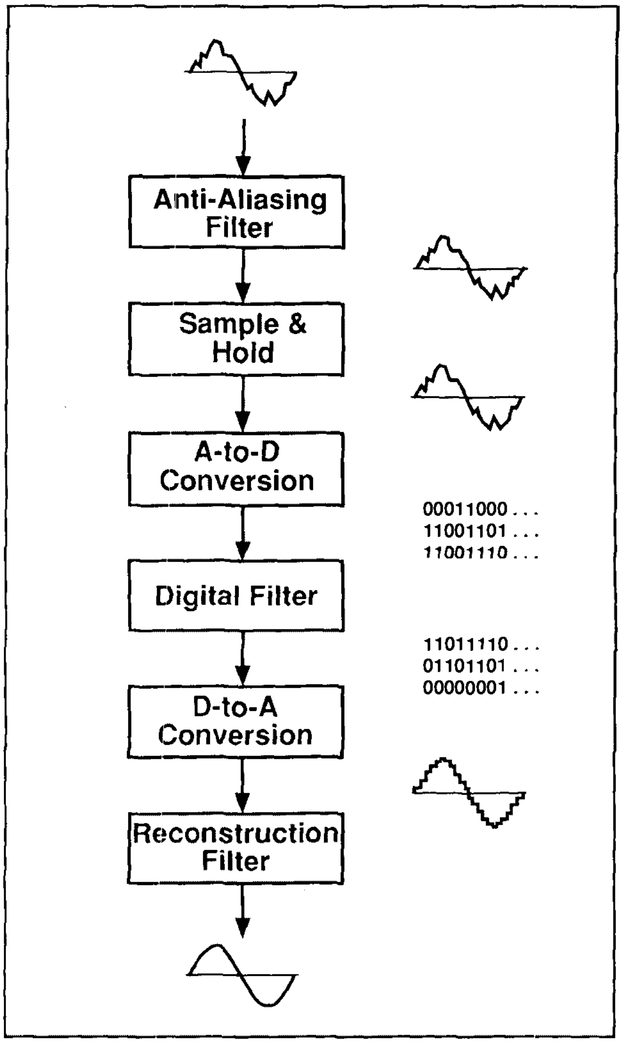
The introduction of transistors and other solid-state devices over the past 20 years has changed the way most of us think about and work with electronic circuitry. As radical as the change from tube-type to solid-state devices and circuits seemed, the current progress in digital signal processing (DSP) technology suggests that an even greater revolution is at hand. Signal processing itself isn't new to electronics or communications — it was used extensively during World War II to remove the noise and clutter from radar signals.[1] Most early signal processing was done in analog form and was of limited value. The introduction of specialized DSP chips has kindled a new interest in signal processing using powerful *digital* techniques. Just as the digital compact disk (which relies on DSP techniques) has redefined the standards in the field of music recording and reproduction, DSP technology promises to drastically enhance the power and flexibility of Amateur and commercial communications systems. This article, the first in a series, examines the fundamentals of digital signal processing and provides experiments in DSP techniques that you can perform on your home computer.

DSP fundamentals

To compare and contrast DSP and classic circuit design techniques, I'll approach a simple problem from both perspectives. For discussion's sake, assume that a low frequency audio signal is accompanied by a significant amount of high frequency noise — for example, a 1000-Hz CW signal accompanied by considerable QRN, most of which is higher than 1000 Hz in frequency at the receiver audio stage. The task is to provide low pass filtering to minimize the amplitude of the high frequency components of the offending noise, without appreciably degrading the desired CW signal.

The standard approach to this problem is to design either an RC or RL low pass filter comprised of discrete components. The DSP approach is considerably more involved and requires several components including an analog antialiasing filter, an analog to digital (A/D) converter, a digital filter and, in the form of hardware and computer instruc-

FIGURE 1



A digital signal processing system configured to provide low-pass filtering of an analog signal.

tions, a digital to analog (D/A) converter and an analog reconstruction filter (see **Figure 1**).

Antialiasing filter

The analog antialiasing filter removes extraneous noise or other higher frequency signals that, when sampled by A/D conversion circuitry, could appear as false low frequency signals. This effect of folding higher frequencies down to lower frequencies is often called aliasing, hence the name antialiasing filter.

A/D converter

The A/D converter can be thought of as a combination sampler, quantizer, and encoder. The A/D converter samples the analog input signal at evenly spaced intervals of time, and produces an accurate, digital representation of the instantaneous signal amplitude.

Of the wide variety of techniques available for analog-to-digital signal conversion, the most common include parallel encoding, successive approximation, voltage-to-frequency conversion, and single slope integration.[2] Each of these basic techniques has its own tradeoffs in terms of relative accuracy, speed, and expense. A/D converters vary in sampling rate, quantization noise, dynamic range, and digitization technique used.

The Nyquist sampling theorem states that the minimum sampling frequency is at least twice that of the signal to be sampled. The sampling rate effectively limits the upper frequency range of signals that can be digitized accurately. To adequately digitize a 50-kHz analog signal, the digitizer should sample the input waveform at least 100,000 times/second. Sampling at a lower rate results in aliasing, a significant source of noise in A/D conversion. One way to avoid aliasing is to use a low pass (antialiasing) filter that cuts out frequencies greater than half of the sampling rate.

Both the quantization noise and the dynamic range of an A/D converter are related to how the digital data is represented. Eight-bit digitizers, which represent amplitude variations with integers from 0 to 255, generally have more quantization noise than 12 or 16-bit digitizers. Quantization noise can be thought of as the high frequency noise introduced into a digitized waveform due to the "staircase effect" inherent in the A/D process. For example, even though the amplitude of an analog signal smoothly varies from 2 to 3 volts, an eight-bit digital representation of the signal might show sudden jumps from 2.0 to 2.5 to 3.0 volts. A 16-bit digitizer, on the other hand, might record less drastic changes. The less jagged representation provided by the 16-bit digitizer should result in less associated quantization noise. You can also think of quantization noise as the difference between the true value and the actual value of a waveform that would be observed on an oscilloscope if the digital representation were immediately applied to a D/A converter.

The digital representation of signal amplitude defines the maximum dynamic range (the difference between the largest and smallest signal that can be recorded) of a digitized signal. The dynamic range of an A/D converter, equivalent to the signal-to-noise ratio of a converter, is related to the number of bits used to represent a waveform by the following equation:[3]

$$SNR = 6.02\ n + \log_{10}\ (1.5)\ dB$$

where n is the number of bits used in the digital representation of the sinusoidal waveform, not including the sign bit. That is, an eight-bit ($N = 8$; $n = 7$) A/D converter uses one bit for sign (+ or −) and seven bits ($n = 7$) for magnitude. For example, with $N = 8$, 12, 16, and 32, the $SNR = 42$, 66, 90, and 187 dB, respectively.

Digital filter

The digital filter (in actuality a computer program or algorithm) works with the digitized signal data to produce the desired output, which is also in digital form. Unlike a conventional system based on discrete components, all waveform manipulations in a digital filter are implemented in software. In many respects, this operation is not unlike everyday data processing. It just so happens that the data represents signal amplitude levels, as opposed to financial account balances or the current value of stocks and bonds. The other qualification is that DSP deals with *real time* analog signals which are processed digitally.

Because of the computational demands of digital filtering, systems based on general purpose microcomputers typically are limited to applications where the signals have bandwidths of only a few hundred hertz. Specialized DSP chips, which can be interfaced to a microcomputer system, extend the working bandwidth by several orders of magnitude. I'll discuss a working example of a microcomputer-based algorithm that implements digital low pass filtering for signals in the audio frequency range later in this article.

D/A converter

The D/A converter performs the inverse function of the A/D converter described earlier. The output of the digital filter is converted by the D/A conversion circuitry into an analog signal. Like A/D converters, D/A converters suffer from the staircase effect, in that a continuous output signal is represented as abrupt changes in amplitude.

Reconstruction filter

Like the antialiasing filter, the reconstruction filter is typically an analog low pass filter composed of discrete components. Its purpose is to remove the high frequency staircase artifact introduced by the D/A conversion process.

Advantages of DSP

Given the above signal processing environment, the digital filter designer would still be faced with the problem of specifying the sample rate for the A/D and D/A converters, selecting the implementation hardware, and creating or locating the appropriate algorithm for the desired characteristics of the digital filter. The designer also has to work around the memory and computational limits of the hardware platform.

With these demands, why would anyone ever consider using DSP techniques instead of conventional methods of circuit design? The major reason is flexibility, especially when complex filtering operations are considered. For simple problems, the DSP solution is much more complex than the conventional alternative. However, this distinction fades when you consider complex filtering involving multiple inputs. While the DSP system would be able to perform all needed operations without a change in physical configuration, a conventional analog system would probably require numerous discrete components for each filtering

operation. A DSP system can be transformed into a specific tool with changes only in programming.

Digital signal processing has additional advantages over analog systems developed with discrete components. DSP techniques are stable, repeatable, and predictable. DSP systems don't degrade with time, suffer from temperature drift, or change operating characteristics with age. The mathematical relationships and binary operations within a digital system remain constant, making the outcome completely predictable and reproducible from one DSP system to another.

Using DSP, the conventional hardware design task is transformed into a programming problem. Once you know the algorithms, it's relatively straightforward to get a process working. The bulk of the development cycle is typically spent improving the speed and performance of the algorithms. Many operations that are difficult or nearly impossible to achieve with conventional analog processing techniques, like linear phase filtering, are almost trivial with today's DSP techniques. DSP promises to minimize the design and testing time inherent in circuit design and maximize the designer's flexibility when solving particular problems.

DSP limitations

DSP techniques are not without their limitations. In addition to quantization noise and limited dynamic range, DSP suffers from round-off and truncation errors, along with limited bandwidth.

Round-off and truncation errors

All digital computations, especially those that use multiple operations on floating point numbers (3.1415926535898 or π, for example) are prone to errors. Because of the way in which the two components of a floating point number (the mantissa and the exponent) are handled, computations involving numbers in floating point format aren't exact. Even when hardware is dedicated to floating point operations, the cumulative results of multiple mathematical manipulations will become significant, given enough iterations.

Round-off errors, like quantization noise and limited dynamic range, are due to computer hardware limitations. They can be minimized by the judicious use of appropriate data types and algorithms. Another type of error, independent of the computer hardware, can be attributed to the algorithms used. *Truncation errors* occur when algorithms use approximations to arrive at an answer. Instead of computing the sum of an infinite series, a practical algorithm might stop after a sufficient number of elements in the series have been added. Truncation error can best be thought of as the difference between the actual answer and the answer obtained via a practical calculation. Unlike round-off errors, which are inherent to the hardware platform, truncation errors can be minimized by selecting the appropriate algorithms for a given problem.

Limited bandwidth

Although there are DSP systems which operate in real time on microwave signals (massively parallel over-the-horizon radar receiving installations[4]), most of us will have to contend with the limited bandwidth of current commercial DSP devices.

The best of today's generally available DSP chips use clock frequencies of up to 42 MHz to realize cycle times of as low as 50 nanoseconds.[3] Depending on how many signals must be handled concurrently, the usable bandwidth of these devices approaches only a few hundred kHz.

DSP hardware

For the past several decades, the use of digital signal processing has been limited by the inherent limitations of available computer hardware. Although digital signal processing can be done with a general purpose computer system, only the relatively recent introduction of special purpose digital hardware has moved DSP from the research laboratory into the hands of engineers and developers. The introduction of VLSI chips that can perform signal processing algorithms in parallel, with many chips operating simultaneously, makes it possible to perform many signal processing tasks up to a thousand times faster than present computers — including supercomputers.[1]

Dedicated integrated circuits

The first generation of dedicated DSP devices, like the INTEL 2920 introduced in 1979, had on-chip scratch pad (RAM) and program memory (EPROM), sample and hold circuitry, and a high speed digital processor, along with D/A and A/D converters.[5] With a 10-MHz clock, this 40-pin DIP chip was fast enough to perform real time processing on analog signals. It was used extensively to implement frequency and phase filters, as well as peak detectors.

Since the introduction of the INTEL 2920, there have been two subsequent generations of dedicated DSP devices. Examples of later generation devices are Texas Instruments' TMS320, Analog Devices' ADSP2100, Motorola's DSP5600, NEC's MPD7720, Fujitsu's MB8764, and AT&T's WE DSP32. These later devices typically provide much more program memory than the 2920, allowing for more conventional programming techniques. That is, some devices provide enough memory to permit programmers to work in high level languages — like C, as opposed to ASSEMBLER.

Compared to the INTEL 2920, which was limited by its on-board A/D and D/A converters, these newer devices have been used to implement a wide variety of digital filters for a variety of applications. The TMS320 has been a popular building block of portable spectrum analyzers.[6] All current DSP chips use a single 5-volt supply, and are designed to interface directly with TTL devices.

Device architectures

DSP chips are designed with one primary consideration: speed of execution. Unlike the serial nature of the general purpose microcomputer, DSP devices use parallel architectures and pipelining to increase execution speed. Unlike the typical microcomputer which handles one instruction at a time, DSP devices perform multiple operations simultaneously. A complex problem that might take 20 instruction cycles to solve in a serial device might take only six instruction cycles in a parallel device.

Most DSP chips make use of the Harvard Architecture, where the program memory is separate from data memory.[3] This allows data to be fetched concurrently with program instructions. The Harvard Architecture contrasts sharply with the architecture of conventional computer systems, in which only a data value *or* a program instruction can be moved from one memory location to another at a time.

In addition to separate data and program memory, DSP chips have a number of dedicated processing units that can work in a highly coordinated fashion. One or more arithmetic logic units (ALUs) handle addition, subtraction, logical operations, and address generation. The shifter scales data to or from other processor units, while the multiplier (MAC) performs multiplications.

DSP chip instruction sets

The instruction sets (an instruction set is the native language of a microprocessor chip) supported by specialized DSP chips are in many respects comparable to those supported by the multipurpose CPUs used in popular microcomputers. DSP devices regularly support an impressive assortment of arithmetic and logic operations, data and program control, condition testing, and system control instructions specifically designed to support signal acquisition and processing.

DSP software

Every DSP program has two components: one that deals with signal acquisition, and one that's concerned with the actual processing. Of the two components, signal acquisition is usually the most straightforward, perhaps because the alternatives are limited. Designers are typically constrained by the resolution, sampling rate, and dynamic range associated with a particular A/D converter, even in a system designed around a general purpose microcomputer. On the other hand, designers have considerable flexibility in the design of the actual signal processing software.

Perhaps the most significant software design issue, especially in a microcomputer-based DSP system, is whether to work in the frequency or time domain — that is, whether to represent signal amplitude as a function of time or frequency. This decision is akin to deciding between using an oscilloscope or a spectrum analyzer as a diagnostic tool; it depends on your experience with the tools in question and on the problem at hand. Traditionally, digital signal processing in the time domain has been avoided in favor of the more powerful processing available when working in the complex frequency domain. Although there's increased activity in investigating time domain signal processing,[9] frequency domain work continues to be the most popular method of signal processing.

Even so, most of us are more comfortable working in the time domain. After all, recording events relative to the passage of time is part of our everyday lives. Many of our test instruments, including the common oscilloscope, assume that time is the point of reference. Electronic signals are commonly expressed as functions of time. Time usually serves as the point of reference in the analysis of common circuits, like the charging of a capacitor. Were you to view the charging of a capacitor on an oscilloscope screen, you'd see a familiar voltage curve increasing exponentially over time.

Anyone who has worked with a spectrum analyzer appreciates the unique perspective it provides. A conventional time-domain oscilloscope simply isn't useful in assessing quantitatively the spectral quality or bandwidth of a signal. Similarly, there are operations in signal processing that just can't be performed easily in the time domain. For example, consider the problem of constructing a linear phase filter — a low pass filter that has linear phase characteristics over the operating frequency range of the filter. The very concept of phase linearity is foreign to the design of high and low pass filters in the time domain, where we work with magnitude characteristics without regard to phase.

By working in the frequency domain with mathematical techniques like the Fourier Transform, a filter designer can manipulate the phase characteristics of a filter easily. Just as a spectrum analyzer is easy to understand and use, designing filters and other circuits in the frequency domain can become second nature once you understand the nature of the device and the mathematical tools involved.

DSP with your personal computer

For signals in the audio spectrum, the capabilities of even the most modest personal computer are sufficient to demonstrate basic DSP techniques. Some microcomputer systems, like the Commodore 64, have hardware dedicated specifically to DSP tasks. I'll describe the DSP capabilities of the C-64 hardware in more detail in the next section, along with two BASIC programs that demonstrate the capabilities of this class of hardware environment.

Audio frequency DSP with the C-64

The 6581 Sound Interface Device (SID) in the C-64 can be used as a simple DSP system. While the 6581 supports several tone oscillators, amplitude modulators, envelope generators, and A/D conversion, the most important characteristic from the perspective of DSP is its programmable filter. The digital filter section of the 6581 can provide high, low, bandpass, and notch filtering, as well as variable resonance and volume control over a range of 30 Hz to 12 kHz.[7] The rolloff is about 12 dB/octave.

The SID chip supports three separate software-definable filters whose outputs can be chained to produce a variety of effects. For example, both low pass and high pass filters can be selected to produce a notch filter response. The volume can be varied from no output to maximum volume in 16 linear steps. Because the SID chip provides no amplification, maximum volume represents no attenuation of the input signal.

The external analog audio input, applied to pin 5 of the C-64's audio/video socket (see **Figure 2**), shouldn't exceed 3 volts peak to peak. The input impedance is on the order of 100 k. The external signal can ride a DC voltage, if necessary, because the external input is AC-coupled to the SID chip through a 1-μF electrolytic capacitor. The audio output of the SID chip, directed to pin 3 of the audio/video socket shown in **Figure 2**, has an impedance of approximately 1 k. The output level, set by the output volume control on the SID chip, reaches a maximum of 2 volts peak to peak. Like the audio input, the audio output of the SID chip is AC-coupled to the audio/video socket through a 1-μF electrolytic capacitor.

Table 1 describes the registers and SID chip addresses pertinent to the discussion of DSP SID offset addresses 21 and 22 define the cutoff frequency used by the filters. Offset address 23 controls both the filter resonance and voice input controls; offset address 24 selects the filter mode and volume.

To better understand how the SID chip can be manipulated for DSP purposes, see the Commodore BASIC program in **Table 2**. This short routine provides an example of digital low pass filtering, with a cutoff frequency of 1000 Hz. The first step is to define the starting address, S or 54272,

TABLE 1

Register Address	Bits	Function
S + 21	□□□□□■■■	Filter cutoff frequency: low-nibble
S + 22	■■■■■■■■	Filter cutoff frequency: high-byte
S + 23	 ■■■■□□□□ □□□□■□□□	Filter resonance/voice input control Filter resonance: 0 - 15 Filter external input: 1 = Yes; 0 = No
S + 24	 □■□□□□□□ □□■□□□□□ □□□■□□□□ □□□□■■■■	Filter mode and volume High-pass mode: 1 = On Band-pass mode: 1 = On Low-pass mode: 1 = On Output volume: 0 - 15

Description of the registers and SID chip addresses pertinent to the discussion of DSP.

of the SID device (line 110) and then to clear the SID registers (lines 140 to 160). Next, define the cutoff frequency (lines 190 to 200). The decimal values to be POKEd into memory locations S+21 and S+22 can be determined by the following formula:

$$S{+}21 = Frequency_{Hz}\ MOD\ 256$$
$$S{+}22 = Frequency_{Hz} - \{(S{+}21) \times 256\}$$

For example, with a cutoff frequency of 1000 Hz:

$$S{+}21 = 1000\ MOD\ 256 = 3$$
$$S{+}22 = 1000 - \{(3) \times 256\} = 1000 - 768 = 232$$

The Commodore, like many other microcomputers, stores multiple part numbers with the low part placed in the first memory location and the high part next. You should note that, because the maximum value that can be POKEd into the low bits of filter frequency cutoff (S+21) is 7 or 111 binary,[8] the cutoff frequency range is from 0 Hz to ((7 ×256) + 255), or 2047 Hz.

Now that you've initialized the SID chip and defined the cutoff frequency, you need to route the external audio input to the filter circuitry (line 220). Line 250 defines the resonance characteristics of the filter. In this example 7, a moderate value of resonance, is selected. To vary the amount of resonance provided by the filter, use the following formula:

$$POKE\ S + 24,\ PEEK\ ((S + 24) + (R \times 16))$$

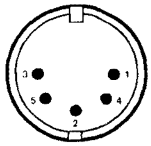
where R is equal to the amount of resonance desired from 0 to 15 (0000 to 1111 binary).

Next, you need to define the output volume of the filter (line 280). Like resonance, output volume can be varied from 0 to a maximum (as in this example) of 15. Remember that maximum volume is really minimum attenuation, as the C-64 doesn't provide amplification of the external audio signal.

Finally, you can define the type of filtering provided by the SID chip — in this case low pass filtering (Line 310). This filter will pass any frequency lower than the cutoff frequency set in addresses S+21 and S+22 without attenuation. See **Figure 3** for a summary of this and other filter mode definitions that can be substituted for line 310 of **Table 2**.

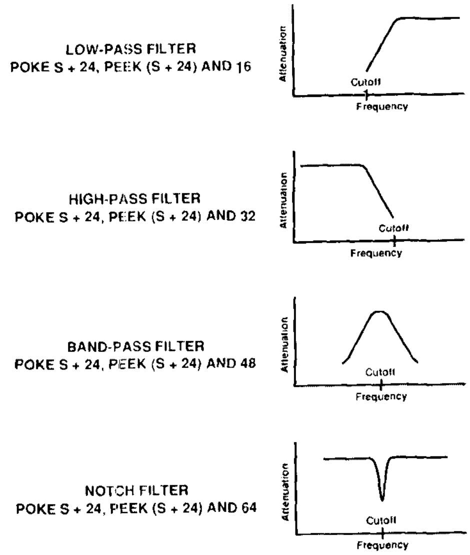
FIGURE 2

Pin	Signal
1	Luminance
2	Ground
3	Audio Out
4	Video Out
5	Audio In



The audio/video port pinouts for the Commodore-64. Pins 2, 3, and 5 are used in the audio frequency DSP program described in the text.

FIGURE 3



Filter mode definitions that can be used with the Commodore-64's SID chip.

Table 3 provides a more interactive DSP environment on the C-64. A standard game paddle can be used to vary the cutoff frequency of a notch filter in real time, from 0 to 2047 Hz. Line 110 defines the memory address for the start of the paddle routine. Because the paddles are unreliable when read from BASIC alone, a machine language routine (lines 340 to 400) is POKEd into memory (line 130). Next, as in **Table 2**, the SID chip address is defined (line 150) and the chip is reset (line 170). The filter is set up to handle an external audio signal (line 190), filter resonance is set to medium (line 210), and filter volume is set to maximum (line 230). Next, the filter mode is set to notch (line 250).

Because the machine language routine to read the paddle values (lines 340 to 400) returns a value between 0 and 255 when called (line 280), the value returned is multiplied by 8 to cover most of the frequency range supported by the filter, 0 Hz to 2047 Hz. For example, when the paddle value is read as 255, the value variable (V) is set to 8 × 255 or 2040. Because the cutoff frequency must be POKEd into the appropriate memory locations as separate high and low order bits, the value variable is decomposed

into V1, the low order three bits, and V2, the high order eight bits (line 290). The high and low order bit values are then POKEd into their corresponding cutoff frequency memory locations (line 300). The endless loop structure set up by line 310 causes the cutoff frequency value to be updated continuously, therefore following the instantaneous game paddle position.

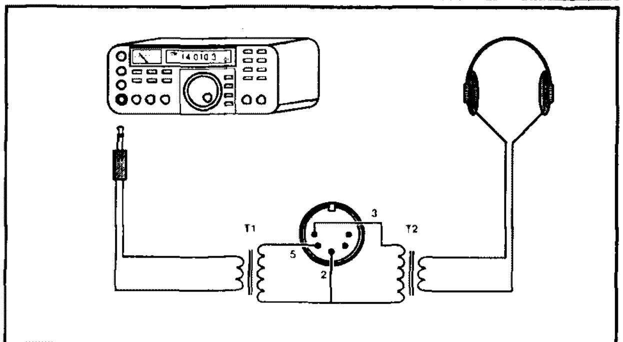
Equipped with the above programs, you can experiment with a variety of audio input and output hardware devices. For example, if you take moderate care to match the input and output impedances of the C-64 hardware with your communications gear, you can experiment with digitally processed CW signals. I found that a pair of common audio transformers, like the popular 1000 to 8-ohm variety (Radio

TABLE 2

```
100  REM DEFINE SID CHIP SOUND ADDRESS
110  S = 54272
120  REM
130  REM CLEAR SID CHIP REGISTERS
140  FOR I = 0 TO 28
150  POKE S + I, 0
160  NEXT I
170  REM
180  REM SET CUTOFF FREQ TO 1KHZ
190  POKE S + 21, 3
200  POKE S + 22, 232
210  REM FILTER EXTERNAL INPUT
220  POKE S + 23, PEEK (S + 23) OR 16
230  REM
240  REM SET RESONANCE TO MEDIUM
250  POKE S + 23, PEEK (S + 23) + (7 * 16)
260  REM
270  REM SET VOLUME TO MAXIMUM
280  POKE S + 24, 15
290  REM
300  REM SET FILTER MODE TO LOW PASS
310  POKE S + 24, PEEK (S + 24) AND 16
```

A BASIC listing for the Commodore-64 that illustrates audio-frequency DSP techniques. This program defines a low-pass digital filter with a cutoff frequency of 1 kHz.

FIGURE 4



A suggested configuration for experimenting with DSP and real-time CW signals. Audio output from a receiver or transceiver is coupled to the relatively high impedance external audio input (pin 5) port of the Commodore-64, through an impedance matching transformer (T1). The digitally filtered audio output (pin 3) is coupled to low impedance headphones through a second impedance matching transformer (T2).

TABLE 3

```
100  REM DEFINE START OF PADDLE ROUTINE
110  C = 12 * 4096
120  REM READ IN PADDLE ML DATA
130  FOR I = 0 TO 63: READ A: POKE C + I, A: NEXT I
140  REM DEFINE SID CHIP SOUND ADDRESS
150  S = 54272
160  REM CLEAR SID CHIP REGISTERS
170  FOR I = 0 TO 28: POKE S + I, 0: NEXT I
180  REM FILTER EXTERNAL INPUT
190  POKE S + 23, PEEK (S + 23) OR 16
200  REM SET RESONANCE TO MEDIUM
210  POKE S + 23, PEEK (S + 23) + (7 * 16)
220  REM SET VOLUME TO MAXIMUM
230  POKE S + 24, 15
240  REM SET FILTER MODE TO NOTCH
250  POKE S + 24, PEEK (S + 24) AND 64
260  REM
270  REM SET CUTOFF FREQ
280  SYSC: V = 8 * PEEK (C + 257)
290  V1 = INT (V/256): V2 = V - V1
300  POKE S + 21, V1: POKE S + 22, V2
310  GOTO 280
320  REM
330  REM  MACHINE LANGUAGE DATA
340  DATA 162, 1, 120, 173, 2, 220, 141, 0, 193
350  DATA 169, 192, 141, 2, 220, 169, 128, 141
360  DATA 0, 220, 160, 128, 234, 136, 16, 252
370  DATA 173, 25, 212, 157, 1, 193, 173, 26
380  DATA 212, 157, 3, 193, 173, 0, 220, 9, 128
390  DATA 141, 5, 193, 169, 64, 202, 16, 222
400  DATA 173, 1, 220, 141, 6, 193, 88, 96
```

A BASIC listing for Commodore-64 that provides an interactive environment for investigating audio DSP techniques. The cutoff (center) frequency of the digital notch filter can be controlled, in real time, by the position of a game paddle attached to the game port of the Commodore-64.

Shack no. 273-1380), are sufficient to couple receiver audio into the C-64 filter system and out to a pair of low impedance headphones (see **Figure** 4). Alternatively, you can experiment with an audio signal generator and your home stereo system or an audio amplifier

Figure 4 shows a suggested configuration for experimenting with DSP and real-time CW signals. Audio output from a receiver or transceiver is coupled to the relatively high impedance external audio output (pin 5) port of the Commodore-64 through an impedance matching transformer (T1). The digitally filtered audio output (pin 3) is coupled to low impedance headphones through a second impedance matching transformer (T2).

Time domain DSP with generic PC hardware

Experimentation with microcomputer-based DSP techniques needn't be limited to computer systems that come factory equipped with DSP hardware, like the Commodore 64, especially if real time processing isn't a necessity. Assuming that there's some way of moving real or simulated signal data into and out of working memory by means of D/A and A/D converter cards or simple data statements written in BASIC, then DSP is simply a matter of programming. (A variety of digital I/O boards and periperals are available for the popular microcomputer systems through a number of sources including MetroByte Corporation, 440 Myles Standish Boulevard, Taunton, Massachusetts 02780.) Although the programs will typically be much more involved than the example listings I've provided for the C-64, they can also be much more powerful. You are no longer constrained to make do with the algorithms provided

TABLE 4

```
100  REM DEFINE VARIABLES, IN/OUT ARRAYS
110  LAST = 0
120  COUNT = 30
130  DIM INPT[COUNT]
140  DIM OUTPT[COUNT]
150  REM
160  REM READ INPUT SIGNAL INTO INPT ARRAY
170  FOR I = 1 TO COUNT: READ INPT[I]: NEXT I
180  REM
190  REM DEFINE FILTER TIME CONSTANT
200  INPUT "Cutoff Frequency (1-100)?", CUTOFF
210  IF CUTOFF < 1 OR CUTOFF > 100 THEN 170
220  X = CUTOFF * 628 / 256
230  ALPHA = (X * X * 35 - 9644 * X + 1000000) / 1000
240  REM
250  REM FILTER INPUT SIGNAL
260  FOR I = 1 TO COUNT
270  TEMP = (1024-ALPHA) * (INPT[I] * 1024)
280  TEMP = (TEMP + (ALPHA * LAST)) / 1024
290  LAST = TEMP
300  OUTPT[I] = INT(TEMP / 1024)
310  NEXT I
320  REM
330  REM PRINT FILTER OUTPUT
340  CLS
350  FOR I = 1 TO COUNT: PRINT OUTPT[I]: NEXT I
360  REM
370  REM STEP SIGNAL OF AMPLITUDE 20
380  DATA 0, 0, 0, 20, 20, 20, 20, 20, 20, 20
390  DATA 20, 20, 20, 20, 20, 20, 20, 20, 20, 20
400  DATA 20, 20, 20, 20, 20, 20, 20, 20, 20, 20
```

A BASICA listing for the IBM-PC and clones that illustrates time-domain DSP with generic computer hardware. The program simulates a first order recursive low-pass filter, with a user-definable cutoff frequency.

in ROM, but are free to develop your own digital signal processing algorithms.

The program in **Table 4**, developed in BASICA for the IBM PC, provides for low pass filtering in the time domain (frequency domain filtering will be considered in a future article). For those who don't have a digital I/O board for their PC, the input data for the filter is supplied by data statements within the program. Remember that in an actual digital processing system, the input and output would be directed, through the I/O board, to a buffer in preparation for processing.

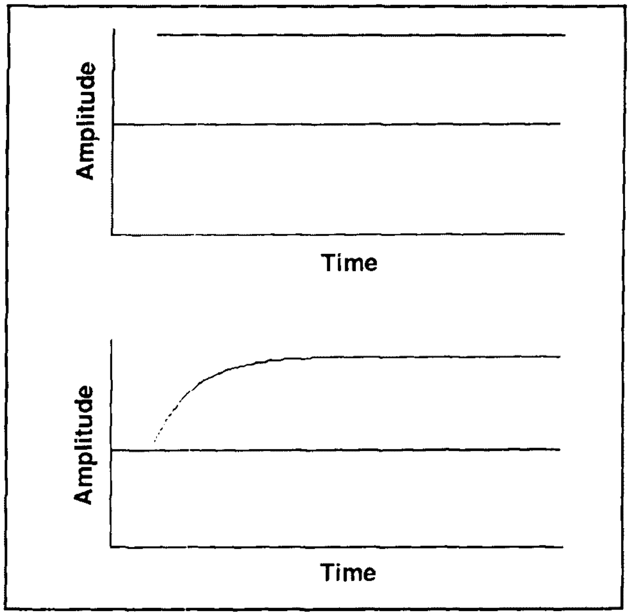
This program simulates a first order *recursive* (IIR) low pass filter, with a user-definable cutoff frequency. The filter is said to be recursive because at any given moment the output of the filter is a function of the current and previous inputs and previous outputs.[10] Recursive filters, while generally superior in performance to nonrecursive filters (filters which don't feed on their own output), suffer from problems associated with instability.[11] While nonrecursive filters are always stable, recursive filters may not be, so the output of a poorly designed recursive filter may increase exponentially — even though the input has ceased. The challenge associated with designing recursive filters comes in selecting filter coefficients that minimize instability while providing the required filter response.

The underlying equation defining the operation of the low pass filter in **Table 4** is:

$$Output_n = \{(K - \alpha) \times (Input_n \times K) + (\alpha \times Output_{n-1})\} / K$$

Where α is the low pass filter time constant, and the coefficient K is taken to be 2^{10}, or 1024. As you can see, the output of this recursive filter is due in part to the previous filter output value ($Output_{n-1}$).

FIGURE 5



Input to (top) and output from (bottom) the first-order recursive low-pass filter defined in *Table 4* (bottom). The step function, defined in the DATA statements of *Table 4*, was sent through the filter with the cutoff frequency set to 1 Hz.

In this example, α is related to the filter cutoff frequency by the following equation:

$$\alpha = \{(Cutoff \times K_1)^2 \times K_2 - K_3 \times (Cutoff \times K_1) + K_4\} / K_5$$

where the coefficients K1, K2, K3, K4, and K5 are defined as 2.45, 35, 9644, 1000000, and 1000, respectively, and Cutoff is the cutoff frequency in hertz. For an excellent introduction to how these coefficients are selected, see the texts by Hamming[12] and Rice.[13]

Given the preceding relationships, it's rather easy to define the filter in a BASIC program. As shown in **Table** 4, the first step is to declare the variables and arrays used in the program (lines 110 to 140). The variable LAST holds the previous output of the filter; that is, $Output_{n-1}$. COUNT is the number of data points in the input signal. In line 170, the input data is read into the input data array, INPT In this example, a step signal of amplitude 20 is used as the input to the filter (lines 380 to 400). Remember that in an actual DSP application, the input data would be routed in real time to the filter input.

Lines 200 to 230 define the filter constant, α in terms of the desired cutoff frequency. Notice that the cutoff frequency range has been limited to between 1 and 100 Hz (line 210). This range will be more than sufficient to examine the effect of digital low pass filtering on the input signal provided. (Note that the cutoff and signal frequencies are relative.)

The actual filter processing takes place in lines 260 to 310. For each input data element, an output value is calculated, based on the filter constant and the previous output value.

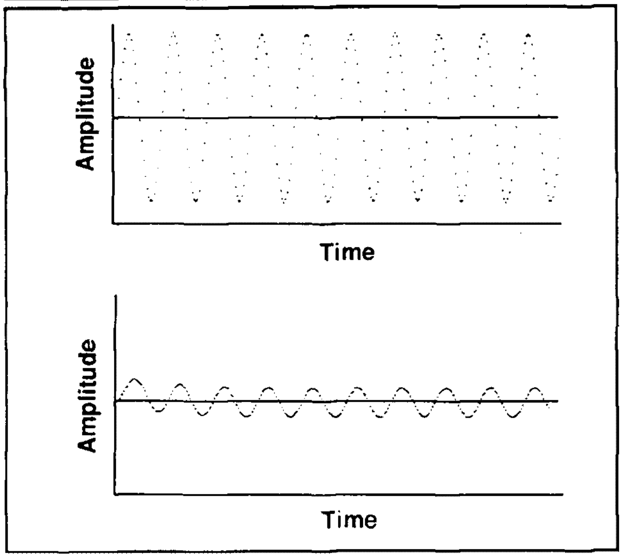
The output of the filter is saved into the LAST variable for the next filter cycle (line 290). In line 300, the filter output is both scaled and converted from a real to an integer value for a more esthetic output. Line 340 clears the screen in preparation for a listing of the filter output values (line 350).

In place of the simple listing of filter output, you can substitute a graphic plotting routine to get a better grasp of the relationship between cutoff frequency and filter output over time. The lower half of **Figure 5** shows the plotted output of the program in **Table 4**, with a cutoff frequency of 1 Hz. To make the plot, the data statement was extended to contain 70 additional input signal points, and the count variable adjusted to 100 accordingly. The filter output was then sent to an Apple Macintosh plotting program for the graphical representation.

The plots shown in **Figure 6** were made in a similar manner by substituting sinusoidal data for the step data in lines 380 to 400 of the program. The input signal (**Figure 6**, top) has a frequency of 10 Hz. Notice the decreased amplitude of the output signal after processing by the digital low pass filter, which has a cutoff frequency of 1 Hz. Upon careful examination of the start of the output waveform (**Figure 6**, bottom), you'll notice that the baseline of the first sine wave is abnormally elevated. The subsequent output waveforms, however, seem to be centered squarely on the baseline.

This aberration is due to the assumptions made during initialization of the LAST variable, corresponding to $Output_0$ in the program. Although an initial value of 0 for $Output_0$ was appropriate for the step function, it wasn't optimum for the sinusoidal signal. If we know a priori about the nature of the input signal (that it always starts with a positive-going sine wave, for instance), then $Output_0$ can be initialized to an appropriate (that is, negative) value, thereby minimizing the baseline distortion.

FIGURE 6



Input to (top) and output from (bottom) the first-order recursive low-pass filter defined in *Table 4* (bottom). To obtain this plot, the DATA statements in *Table 4* were modified to represent the 10 Hz sine wave with amplitude 40. As in *Figure 5*, the cutoff frequency of the filter was defined as 1 Hz. Note the elevated baseline of the first sine wave processed by the filter.

Modifying a prototype low pass filter made of discrete components to create a high pass filter would likely entail several minutes of additional soldering, in addition to the time it would take locate components with the appropriate values. Because you're working with algorithms and not components in the DSP domain, circuit modifications are mere keystrokes away. For example, the low pass filter described in **Table 4** can be modified to work as a recursive high pass filter if you substitute the following line for the one in the complete listing:

```
270 TEMP = 1024 * ((INPT[I] * 1024) - LAST)
```

This BASIC statement, together with the supporting code in lines 260 to 300, implements the following relationship:

$$Output_n = \{(K\ ((Input_n \times K) - Output_{n-1})) + (\overset{8}{a} \times Output_{n-1})\}\ /\ K$$

where K and $\overset{8}{a}$ are defined as in the original low pass filter design.

Summary

DSP may one day make conventional circuit design obsolete. Designers will no longer be concerned with component values and tolerances, but will deal with algorithms and software. In part 2, I'll consider techniques used for digital signal processing in the frequency domain, including the Fourier Transform. *hr*

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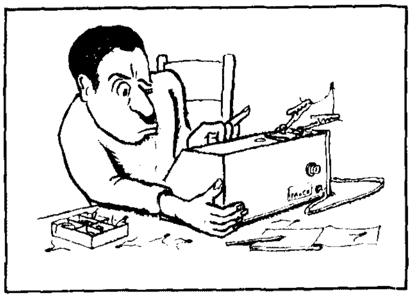
Practically Speaking



By Joseph J. Carr, K4IPV

SOME ANALOG METER APPLICATIONS

In March I looked at analog meter movements, mostly those of DC meters. This month I'll expand the discussion to include some application issues (including meter movement protection) and a construction project.

Meter protection

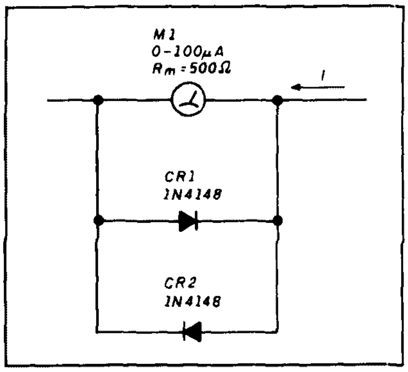
The electrical environment for meters is hostile enough that meter movements seem to be perpetually at risk of total destruction. There can be physical damage. A bent pointer can result when the meter is connected into a circuit where a severe overrange current flows, or if the meter is connected backwards in the circuit. The meter can also fail to perform properly because of the strong RF fields inside a piece of ham radio transmitting equipment. Fortunately, there are a few things you can do to protect your meter.

Overcurrent protection

The simplest way to prevent meter movement damage from overcurrent is to place a fuse in series with the meter. This approach is reasonable when the meter has a current range compatible with fuse values, but there's no fuse for 1-mA (or less) meters. In fact, meter movements much under 100 mA probably can't be protected easily with a fuse.

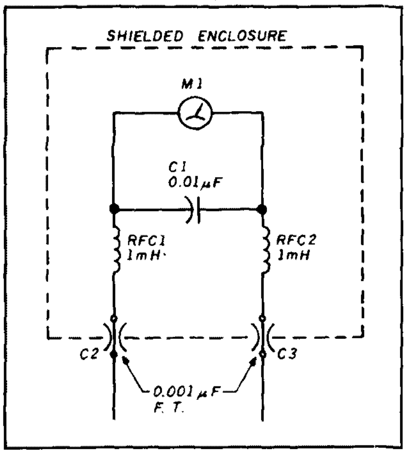
Figure 1 shows a method used in some multimeters to prevent meter movement destruction when an overrange current tries to flow in the movement. Diodes CR1 and CR2 are silicon diodes (e.g., 1N4148) of the type used in low current meters; rectifier diodes (e.g., 1N400x) are used in high current meters. The silicon diode requires a forward bias potential of 0.6 volts or more for proper linear operation, and is virtually ineffective at potentials lower than about 0.3 volts. The voltage drop across the meter is the product of the current flowing in the meter and the meter's internal resistance (R_m). In **Figure 1**, the meter movement is 100 μA full scale and has an internal resistance of 500 ohms. (Always check; the value in your meter will probably be different.) The full-scale current will produce a voltage drop of 500 ohms × 0.000100 A, or 0.05 volts. Under normal circumstances, the voltage drop won't forward bias CR1 and CR2. However, when a severely excessive current flows, a larger voltage drop will occur and the diodes will become forward biased. When this happens the diodes shunt the meter, protecting it from harm (though probably destroying themselves).

FIGURE 1



Protection diodes shunted across the meter movement.

FIGURE 2



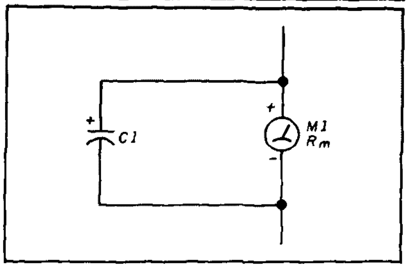
RF protection diodes.

RF protection

RF protection is a must for meters used in transmitters. **Figure 2** shows several of the basic methods used to protect analog meters in transmitters and linear amplifiers. It's always wise to shield the meter in the RF environment. Either use an aluminum box or fashion a shield of your own design from sheet metal. The meter movement is usually mounted directly to the front panel of the equipment, so the shield can also be mounted in that fashion. This simplifies the design and makes simple boxes a reasonable approach.

In most cases, you can get away with using a single bypass capacitor across the meter terminals. Use a disk ceramic capacitor (or other type that will work at the frequencies used in the equipment) between 0.01 μF and 0.5 μF. Make sure the capacitor is mounted directly on the body of the meter movement, stretched between the electrical connection terminals of the meter. The capacitor leads should be as short as possible. In some cases, especially with high powered amplifiers, the RF chokes (at least 1 mH) should also be added in series with the meter movement. Capacitors C1 and C2 are 0.001-μF feedthrough capacitors that are mounted directly to the shielded metal enclosure. If these aren't available or practical, use a 0.01-μF disk ceramic capacitor mounted directly on the input connections to the shielded housing. In any event, regardless of how many of these fixes are incorporated into your project, keep all leads inside the shielded

FIGURE 3



Integrating or low-pass filtering the meter movement.

enclosure as short as possible to minimize RF pickup.

Damping

Some signals have either too much variation or too much noise present on them to be read effectively on an analog meter movement. In those cases it's possible to *integrate* (time average) the meter reading by connecting a capacitor across the meter terminals (see **Figure 3**). This is an old trick in the instrumentation trade. I've seen capacitors from 0.1 μF to 500 μF strapped across the meter terminals. The time constant of the circuit (R_mC1) tells you something about the amount of averaging in any given case.

Amplified meters

The bare meter movement is sometimes boosted with an amplifier circuit to improve sensitivity or perform certain specific instrumentation jobs. One application (see **Figure 4A**) is to provide a higher input impedance to a microammeter or milliammeter (M1) that's connected as a voltmeter. The sensitivity of a 0 to 1-mA or even 0 to 100-μA meter movement might be too poor for practical use, especially in solid-state circuits. But if the meter is *buffered* with an operational amplifier (op amp), as in **Figure 4A**, then the circuit sees the high input impedance of the op amp (A1). In order to maintain that high impedance, be sure to use a device with a very high input impedance. The CA-3140 type is a good example of such a device, and it's both inexpensive and widely available from mail order and local parts sources.

In the circuit of **Figure 4A** the op amp (A1) is connected in the *unity gain noninverting follower* configuration. When a DC voltage (V_{in}) is applied to the +IN terminal of the op amp, exactly the same potential appears at the output terminal; that is, point A. The full-scale input voltage and the series resistors are used to custom tailor the meter movement for voltmeter service. The sum of R1 and R2, denoted R, is found from:

$$R = \frac{V_{in\,(max)} - I_{fs}R_m}{I_{fs}} \qquad \textbf{(1)}$$

where R is the sum R1 + R2, $V_{in(max)}$ is the maximum allowable full-scale input voltage, I_{fs} is the full-scale current of meter M1, and R_m is the internal resistance of M1.

If you want the meter to see an amplified voltage, then use the alternate *noninverting follower with gain* circuit[1] of **Figure 4B**. The DC voltage gain of this circuit will be:

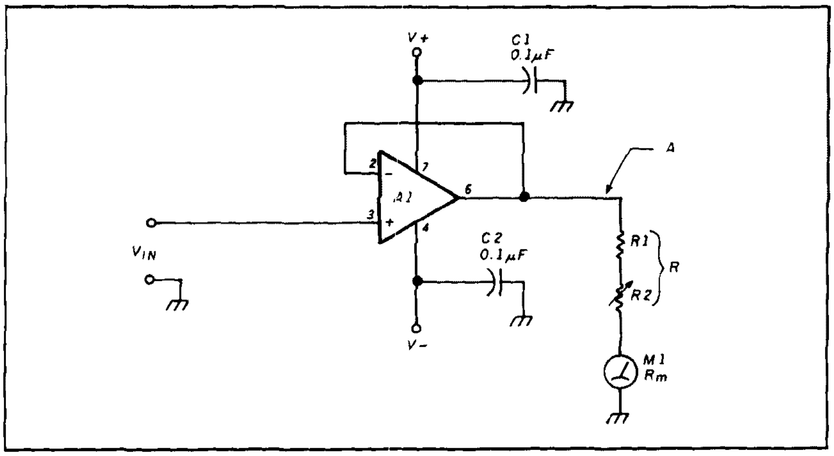
$$A_V = \frac{R4}{R3} + 1 \qquad \textbf{(2)}$$

The voltage at point A (**Figure 4A**) will be A_VV_{in}. In that case, the *maximum* voltage at point A will be used in place of $V_{in(max)}$ in **Equation** 1 to find R.

The reason for making the meter multiplier resistor from two resistors (R1 and R2) instead of a *single resistor* is twofold. First, it allows a series combination of a fixed resistor and a potentiometer to replace a hard-to-find precision resistor. Second, it lets you adjust the meter to full scale easily. Apply the maximum value of input voltage to the noninverting input of the op amp and then adjust R2 for a full-scale deflection of M1.

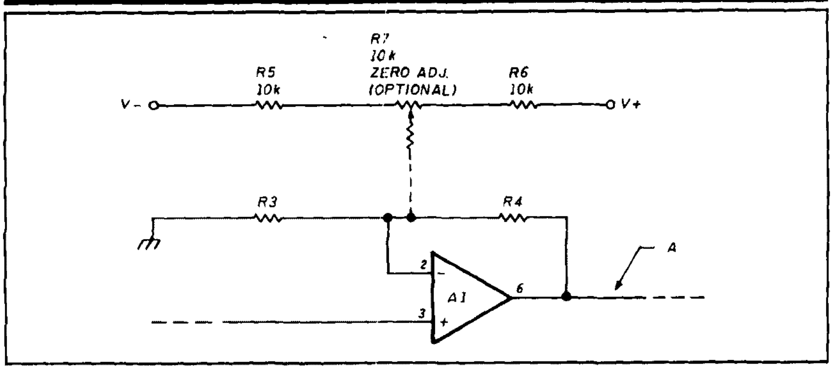
Figure 4B shows an optional **zero adjust** circuit. This circuit injects a counter current into the *summing junction* of the op amp (at the inverting input). This current is used to counteract any DC offset that's present in the op amp output. To adjust R7, short the input terminal (the noninverting input of A1) to ground, and mechanically zero the meter movement pointer

FIGURE 4A



Unity gain voltmeter.

FIGURE 4B



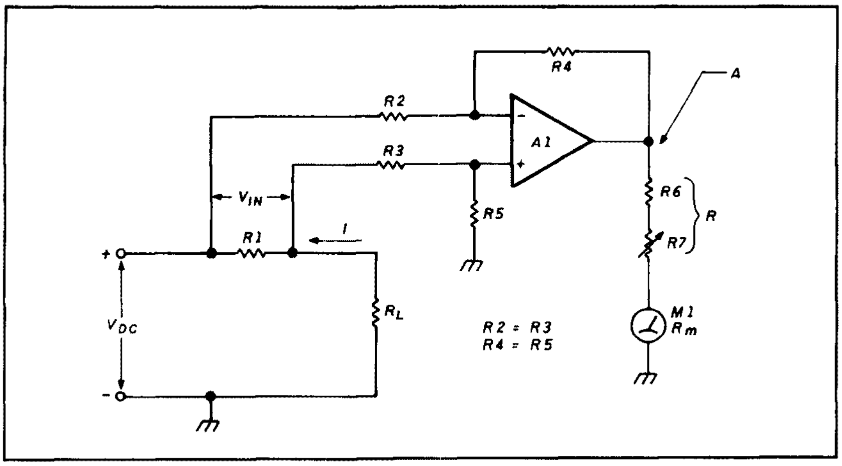
Gain voltmeter.

(there's usually a small screw adjustment on the front panel of the meter). Next, turn on the circuit and adjust R7 for exactly zero reading on the meter.

Figure 5 shows how to make a milliammeter or microammeter into a higher current instrument with an amplifier. Resistor R1 is a small value, precision resistor that's placed in series with the load (R_L) of the circuit. The current flowing in the circuit without this resistor is V_{dc}/R_L, and with the resistor it reduces slightly to $V_{dc}/(R_L + R1)$. It's very important to make $R1 << R_L$ to prevent errors.

The op amp is connected in the *DC differential amplifier* configuration, and has a differential DC voltage gain of R4/R2. The input signal seen by the amplifier (A1) is the voltage drop across R1, which is $I \times R1$. The voltage appearing across the meter circuit at point A is $V_{in}R4/R2$. It's important to keep R2 = R3 and R4 = R5. The values of R6 and R7 are found in the same manner as in the example of **Figure 4A**.

FIGURE 5



Amplified current meter.

Precise AC voltmeters

The AC voltmeter is made from the DC meter movement plus a rectifier. Last month you saw a bridge rectifier that can be used to make the meter read AC values. Unfortunately, that circuit is limited when it comes to very low voltage circuits. The problem is caused by the fact that PN junctions are diodes which are used to rectify the AC before applying it to the meter movement, and aren't linear in the low range of forward bias voltage (see **Figure 6A**). The ideal diode I-versus-V curve is a straight line from zero to maximum current. It's said to be an *ohmic device* because it obeys Ohm's law. But actual diodes are ohmic only in the region above a certain critical voltage (V_g); this voltage is 0.2 to 0.3 volts in germanium diodes and 0.6 to 0.7 volts in silicon diodes. Below V_g the diode is nonlinear, and therein lies the problem.

A "fix" for nonlinearity is to replace the diode with a *precise rectifier circuit* like **Figure 6B**. This circuit uses an op amp to "servo-out" the nonlinearities and make the diode act more like the ideal diode. That's why this circuit is also called the *ideal rectifier* or *ideal diode* circuit. Diodes CR1 and CR2 are ordinary silicon signal diodes like 1N914s and 1N4148s.

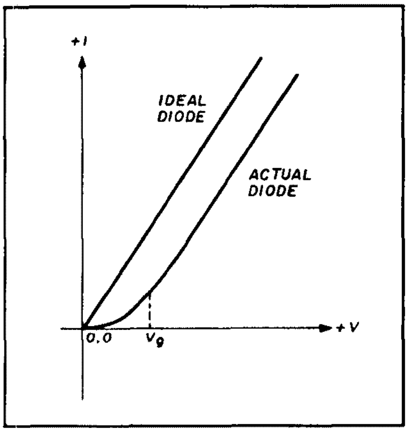
The diodes are shown in **Figure 6B** in the direction that will produce a positive V_o with positive V_{in}; this circuit is called a *positive precise rectifier*. Reversing the diode direction (turning them around) will produce a positive output voltage with negative input voltage (*negative precise rectifier*).

Unfortunately, the circuit of **Figure 6B** is only a half wave and won't prove as useful in many applications. A *full-wave precise rectifier*, also called an *absolute value circuit*, is shown in **Figure 6C**. This circuit combines a positive precise rectifier and a negative precise rectifier into a single output. The output is obtained by amplifier A1, which is a gain-of-one summer. As long as all three resistors have the same value, the output voltage (V_{odc}) will be the sum of the two rectifier outputs (V_{01} and V_{02}), and will be a full-wave rectified pulsating DC waveform.

RMS AC voltage measurement

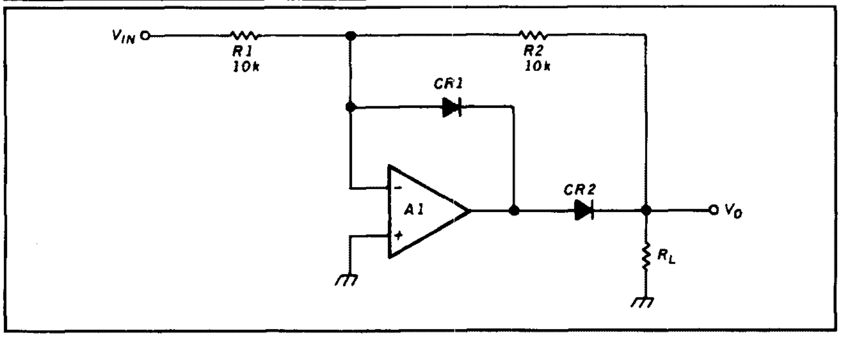
Simple rectified meters produce a reading that's related to the average voltage of the pulsating DC waveform produced by the rectifier. It can be interpreted to be the *root mean square* (rms) voltage only if the scale is so calibrated *and* the input waveform remains a sine wave. However, that condition isn't often met in practice. A

FIGURE 6A



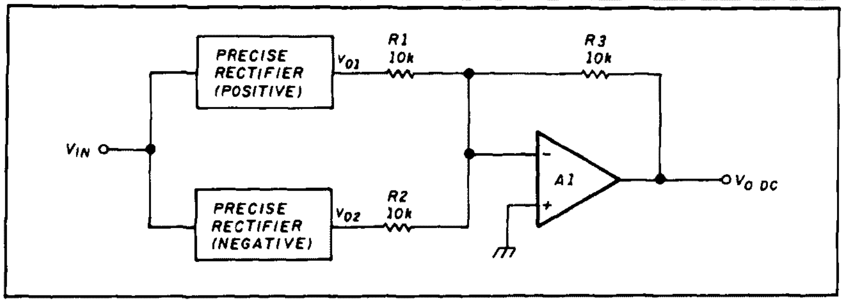
Ideal and actual diode I-versus-V curves.

FIGURE 6B



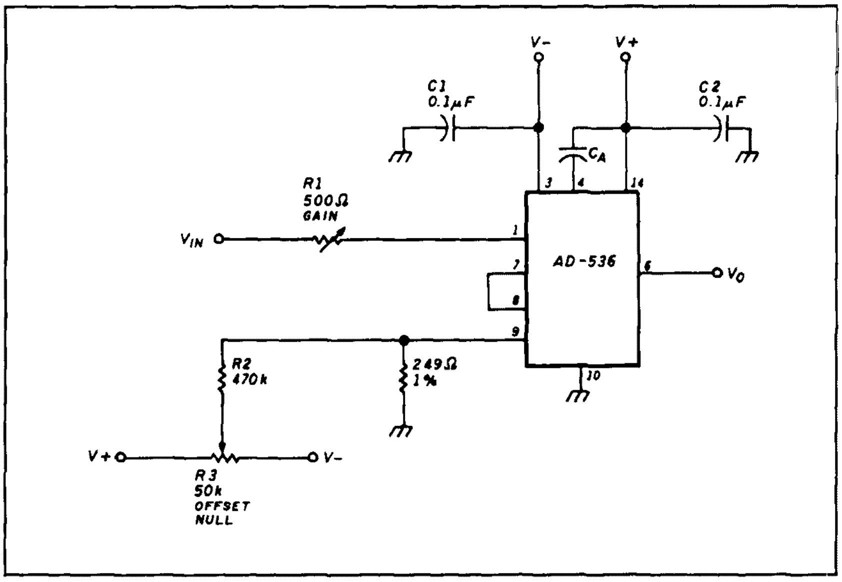
Half-wave precise rectifiers.

FIGURE 6C



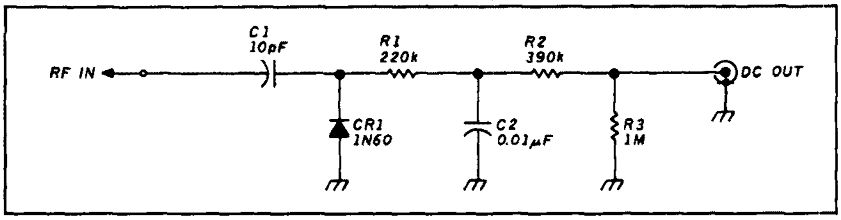
Block diagram of a full-wave precise rectifier.

FIGURE 7



RMS-to-DC converter.

FIGURE 8



RF demodulator probe.

true rms voltmeter requires an *rms-to-DC converter* circuit. **Figure 7** shows an rms converter based on the Analog Devices, Inc. AD-536 rms-to-DC converter. The output voltage will be a DC voltage that's proportional to the rms value of the applied AC voltage. The AD-536 will operate properly at frequencies to 300 kHz if the input rms voltage is >100 mV, and 2 MHz if the input rms voltage is >1 volt. This circuit offers *offset null* (zero) and *gain* controls.

RF voltmeters

Many Amateur Radio applications require measurement of RF voltages. For example, the output power of a CW transmitter can be determined in many cases by measuring the voltage across a dummy load. In other cases, troubleshooting requires such measurements. **Figure 8** shows a simple, passive RF demodulator probe. This circuit *must* be built *into* a shielded enclosure, and consists of a rectifier (CR1) and an RC ripple filter that uses values consistent with the RF frequencies. The diode used should be a germanium diode of the 1N34, 1N60, or ECG-109 type (also NTE-109). This circuit works for voltages up to about 40 volts peak. Increased voltages are obtained by series connecting two or more rectifiers.

Bridge amplifier: a construction project

There are many applications for a differential voltmeter, especially in circuits like the Wheatstone bridge used for measuring resistive impedances and other parameters. The circuit in **Figure 9** can be used for either single-ended or differential applications. **Photo A** shows the completed project.

The input or "front end" of the differential voltmeter is an instrumentation amplifier consisting of three operational amplifiers: A1, A2, and A3. The two input amplifiers, A1 and A2, should be RCA/GE CA-3140 devices or other high input-impedance equivalents. These devices are easily obtained on the mail order parts market. The gain of this section of the amplifier is found from:

$$A_v = \frac{2\,R2}{R1} + 1 \qquad \textbf{(3)}$$

For the value shown, the differential voltage gain of this section is ×20. The presumption is that R2 = R3.

The follow-on post amplifier (A3) has a gain of:

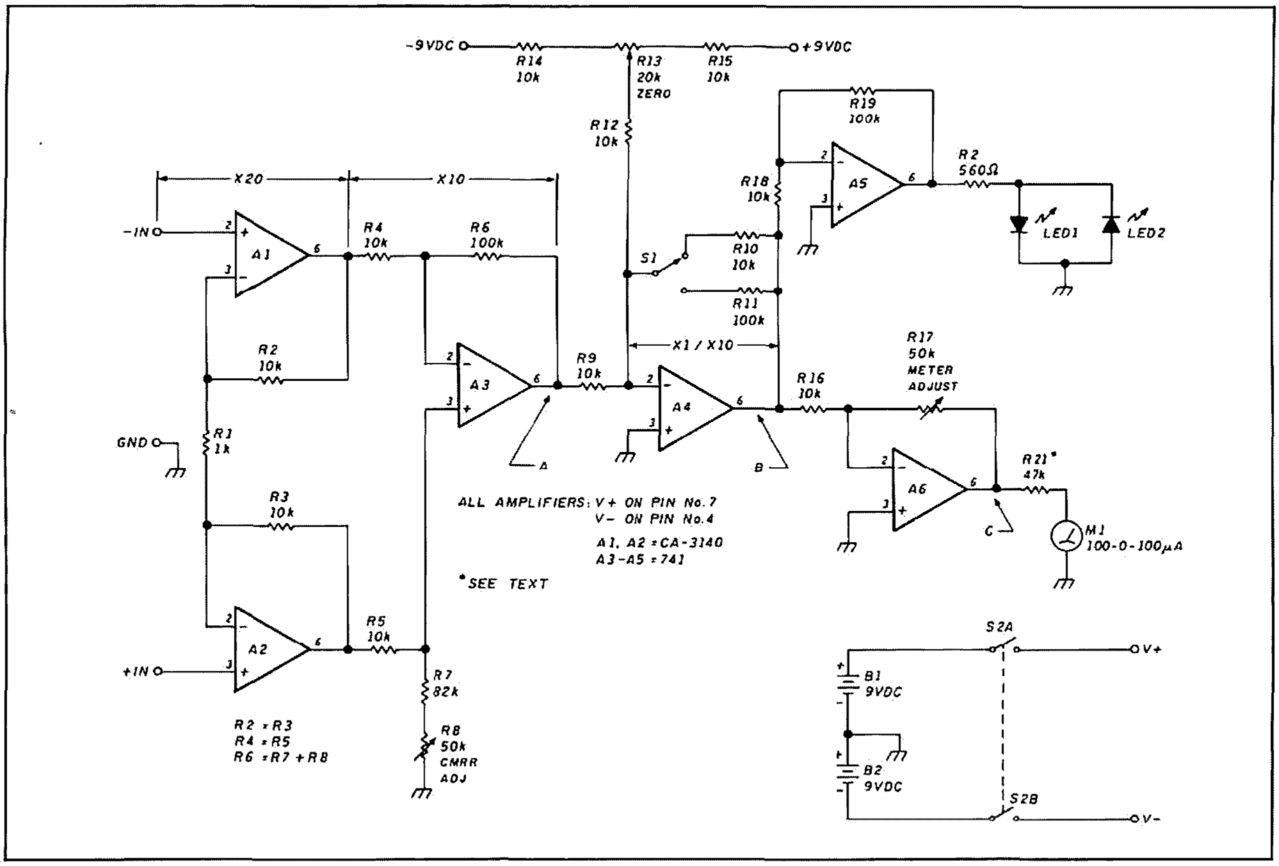
$$A_v = \frac{R6}{R4} \qquad \textbf{(4)}$$

assuming that R4 = R5, and R6 = R7+R8. With the values shown, the gain of the post amplifier is ×10, so the overall gain of the instrumentation amplifier section is the product of the two gains, or ×200.

The *scaling amplifier* (A4) has a switch selectable gain of ×1 or ×10, making the overall gain of the circuit either ×200 or ×2000. With this higher gain, the DC offsets of the amplifiers can be quite high, so a ZERO control (R13) is provided.

There are two outputs on the

FIGURE 9



Differential voltmeter circuit.

differential voltmeter. The conventional output uses a 100 to 0 to 100 μA DC meter in the manner described previously. The second output stage is a null detector made from light emitting diodes (LED1 and LED2). The null condition is indicated when **both** LEDs are extinguished.

Initially, you should adjust the meter with the controls in the following positions:

−IN/+IN: shorted to ground
S1: in the ×10 position (to R11)
S2: off
R8, R13, R17: midrange

Turn on the meter power switch (S2) and let the instrument warm up for about ten minutes. If the meter movement is pegged, adjust R17 to bring it back to the proper range. Adjust CMRR ADJ (R8) for zero volts at point A. Next, adjust ZERO (R13) for zero volts at point B. Now, unground −IN/+IN and connect these two terminals together. Apply a 1.5-volts DC signal from a dry cell to the shorted inputs. Readjust R8 for zero volts at point A if any change has occurred.

PHOTO A

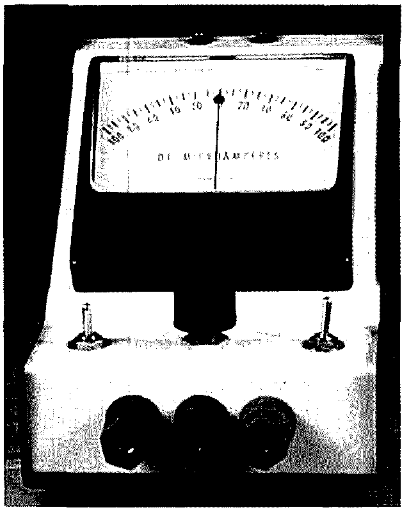


Photo of *Figure* 9 project.

Recheck the voltage at point B and readjust R13 if necessary.

If a precision value of V_{in} is to be read, then apply a 1-mV DC potential to −IN, and ground +IN. Adjust R17 for the proper deflection of M1. This mode isn't normally needed, however, because the purpose of the meter is to find a null.

The DC differential voltmeter is usable as is, or you can use the information contained in this article to design your own application. Don't try to obtain gains higher than ×2000, because the drift and other defects will conspire to cause problems. You need to use special compensation methods in those cases.

Conclusion

The analog meter movement can be used in a variety of applications. These are some of the more popular applications of those meters used in conjunction with other circuitry. hr

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INSTRUMENTATION SERIES: OSCILLOSCOPES

Denton Bramwell, K7OWJ, Director Heath Instruments, Heath Company, Benton Harbor, Michigan 49022

If you like to tinker and build, or if you just want to make routine equipment repairs, an assortment of electronic test equipment is undoubtedly on your "must own" list. If you're like me, you have a pretty strong attachment to your test equipment. After all, your equipment lets you put your knowledge to work.

Unfortunately, not all test equipment is created equal — even equipment with exactly the same specs. I'd like to talk about the pros and cons of various pieces of apparatus to help you select the right tools for your purposes, and get the most out of them once you own them. My topic for this article is oscilloscopes.

Years ago there was a question on the General class exam that asked which measurement tool was the most useful of all. The correct answer, to no one's amazement, was the oscilloscope. It's still a good answer. If you could own just one piece of equipment, you'd probably be wise to choose a good 'scope because it lets you look at both static and dynamic voltages.[1] Choosing a good 'scope isn't difficult if you take the time to learn a few indicators.

Note that I said a "good" 'scope. There are a lot of junky ones on the market, and your first task is to sort out the equipment that's part of the solution from the equipment that's part of the problem. A discussion of what constitutes a good 'scope can be broken down into sections: vertical amplifiers, timebases, trigger circuits, and extra or convenience features. Of these, the vertical amplifier portion of the oscilloscope deserves the most discussion.

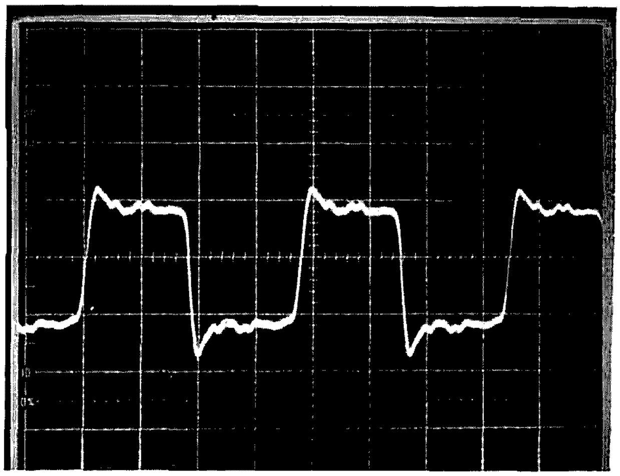
Vertical amplifiers

Look at almost any oscilloscope ad and you'll find the bandwidth specification displayed prominently. In fact, for many buyers bandwidth has too often become the single figure of merit for oscilloscopes. That's too bad; buyers who consider only bandwidth when buying an instrument are courting disappointment. To further compound the problem, a lot of people misunderstand and are misguided by the meaning of bandwidth.

In a nutshell, you probably need a lot more bandwidth than you think you do. Looking at your 9-MHz IF? A 10-MHz 'scope ought to do the job with some room to spare, right? Well, no. In fact, not by a long shot.

Photo A shows the output of an oscillator designed to what I'd call minimal standards. The oscillator was made from a fast logic gate and crystal, breadboarded with little regard for layout. The result is a 13.3-MHz signal, which could charitably be described as "rich in harmonics." What you see in Photo A is a reasonably accurate representation of the output of the circuit taken from a 200-MHz oscilloscope.

PHOTO A

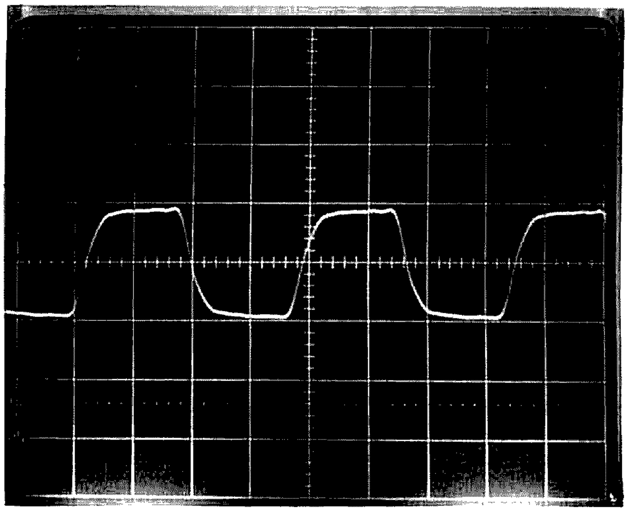


A 13.3-MHz square wave, rich in harmonics, as seen on a 200-MHz oscilloscope.

Now look at what happens when you try to use an instrument with inadequate bandwidth. Photo B shows the same signal, with the same sensitivity and sweep settings, on a 40-MHz oscilloscope. You'd think that a 40-MHz instrument would be ample for a 13.3-MHz signal, but in this case it's not. Photo C shows a worse case, from a 20-MHz oscilloscope. The bandwidth of the instrument is 50 percent higher than that of the signal, but the on-screen trace is nothing like the one from the 200-MHz 'scope. Why do you need so much bandwidth for a 13.3-MHz signal? The reason is easily understood.

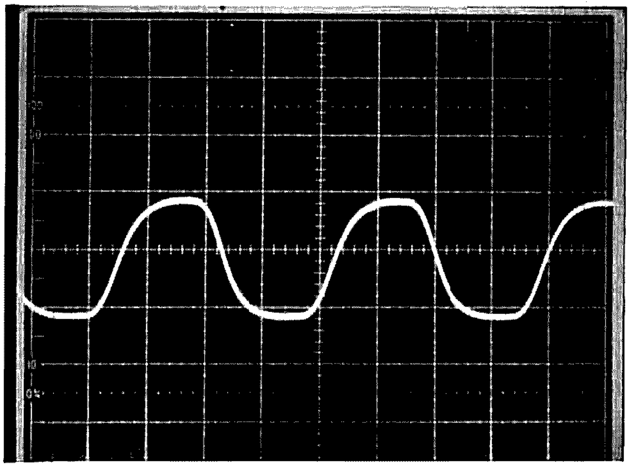
Bandwidth is *not* the same as frequency range. When some instruments (like frequency counters) are specified, the term frequency range is used because counters work properly up to some limit and then cease counting. Oscilloscopes don't work up to a particular frequency and stop. They start rolling off slowly until they reach the point where they are 3 dB down. We call that the bandwidth. In short, a pure sine wave will appear at 70.7 percent of its true amplitude if its frequency is equal to the bandwidth of the 'scope.

PHOTO B



Same 13.3-MHz signal, as seen on a 40-MHz oscilloscope.

PHOTO C



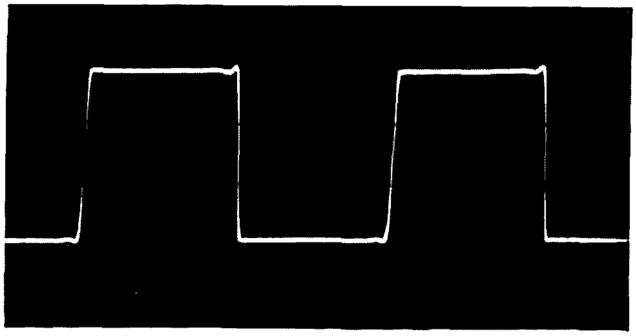
Same signal again, but on a 20-mHz oscilloscope.

If you know in advance that you're dealing with a pure sine wave, you might find such a measurement useful — but this is seldom the case. In fact, if you know that your signal is only a pure sine wave, you can get all the information you need from an RF voltmeter and a frequency meter. Normally circumstances don't justify the assumption that your signals are sine waves. Actually, in RF work it's often the deviations from sine waves that are interesting.

If you don't know in advance what your waveforms look like, you need a bandwidth *much* greater than your signal. An ideal square wave contains an infinite series of odd harmonics. To obtain a reasonable representation of a square wave, you probably ought to think about getting an instrument with a bandwidth five to seven times the frequency of the signals you want to look at. And next time someone tries to tell you that his 35-MHz oscilloscope is really good

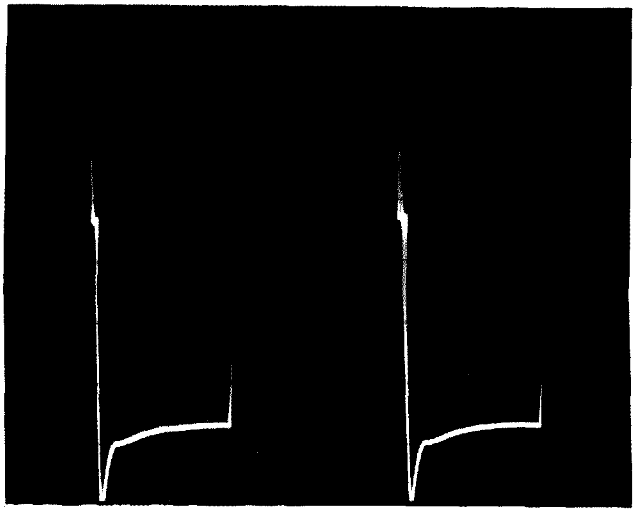
✓ 148

PHOTO D



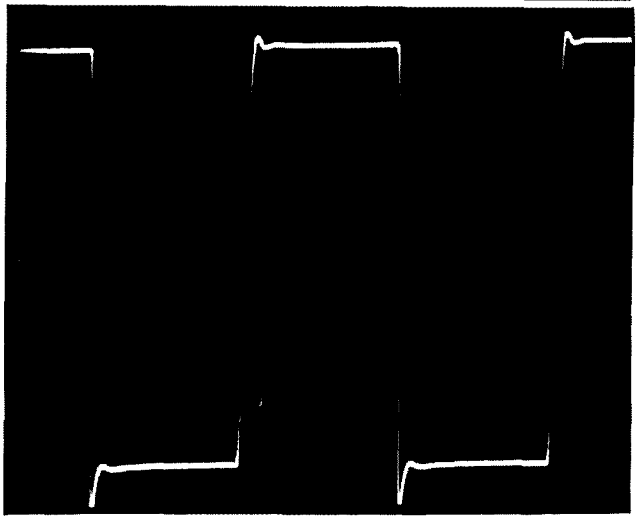
High quality square wave.

PHOTO E



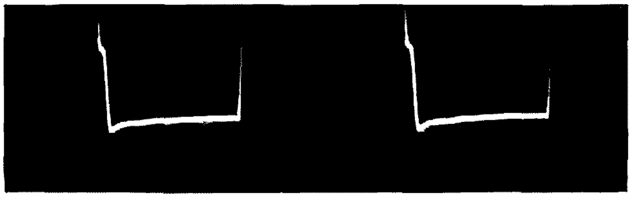
Same square wave, same oscilloscope, but at twice the sensitivity.

PHOTO F



Same square wave again, but at a sensitivity that puts half the signal offscreen.

PHOTO G



Same signal, same setttings, except that the vertical position control has been changed.

to 40 or 50 MHz, you'll know better! Even for sine waves, you'll want a bandwidth three to five times greater than your signal. Otherwise, your measurements will be in error by more than a reasonable factor. A 35 to 65-MHz instrument, rather than a 10-MHz instrument, is what you want for measurements on your 9-MHz IF.

Once you've decided on a bandwidth, you need to look at the pulse characteristics of your instrument. This is necessary *even if your're not looking at pulses.*

The impulse response of a system is the Fourier transform of its transfer function. This is simply a $64 way of saying that if you tell me exactly how something responds to a very fast pulse, I'll tell you exactly how it responds to *anything.* In other words, if you mess with pulse response you also mess with bandwidth. The consequence of accepting poor pulse response is that you also accept sensitivity that isn't flat and doesn't roll off in a neat, orderly manner. You'll find bumps and dips in the sensitivity of the instrument that are well within its useful range. So what do you look for, and how good is good?

The easiest spec of pulse fidelity is the overshoot and ringing spec. A small percentage of overshoot and ringing extends the bandwidth of the oscilloscope greatly. The usual practice is to allow 3 to 5 percent. Be very suspicious of any 'scope that doesn't have this spec. Some manufacturers ignore this parameter altogether. It lets them offer a very high bandwidth for a very low cost, but you won't be satisfied with the instrument for long.

It's important to look for a combination of a good bandwidth spec and good pulse response when you're checking out a new oscilloscope. However, there's one other parameter that's seldom mentioned, and it's one that's very important. I'm talking about dynamic range.

Photo D is taken from a 'scope that's a bit short on this parameter. It shows a square wave from a good quality signal generator. The representation is pretty good. There's a little negative tip on the negative-going slope that really isn't there, but it's small and well within reasonable specs. Now double the sensitivity and use almost the entire screen. You'll see that those innocuous little tips have grown to problem size in Photo E. Advance the gain a little more, as in Photo F, and they really get unruly. Position the trace upward, as shown in Photo G, and the picture changes dramatically once again.

You should expect *some* errors of this type; however, these are a bit large. Unfortunately, I've seen products far worse than this. Since dynamic range is seldom specified, it's an item you'll probably have to check for yourself using the method shown here.

Here's another very common trap. You plunk down $700 for a racy new 50-MHz oscilloscope, and then spend $50 on a pair of 50-MHz probes. Sounds reasonable — a nice, matched system. It's not. A 50-MHz probe on a 50-MHz 'scope will give you about 35 MHz at the probe tip! And the degradation is even more severe when you hook it to a point in your circuit with more than 50 ohms of source impedance!

It's better to spend a few more bucks for probes of 250 MHz (three to five times your 'scope bandwidth) or so. That will preserve almost all of your bandwidth at the probe tip, and give you more megahertz per dollar.

One more thing before I go on. Check out the cabinet material before you buy. You'd be surprised how many plastic 'scopes suddenly get inch-wide traces when you key your transmitter. A metal case doesn't guarantee immunity, but it does better the odds. I've found that plastic-cased oscilloscopes tend to be more EMI susceptible, and should be avoided unless they have good internal shielding.

There's a lot of information out there about vertical amplifier sections,[2] but what I've discussed here should provide you with a better than average survival kit. These simple questions and performance tests will keep you from making the most common errors in oscilloscope selection and use.

Timebases

Old, low cost oscilloscopes used a horizontal oscillator with a synchronizing circuit. The idea was to synchronize the frequency of the timebase with the frequency of your signal, thus obtaining a usable display. That style of design has pretty much become a thing of the past.

The calibrated and triggered timebase has played a major part in making oscilloscopes as popular and useful as they are today. A timebase is fairly easy to check out; it's supposed to be linear. Just feed in any signal known to produce peaks or zero crossings at a rate of one or five times per division, and look for errors. If the largest error on any timebase setting is 2 to 3 percent, the 'scope is doing as well as analog models generally do. Photo H shows a typical example.

Usually, the oscilloscope has a magnifier which allows you to expand the trace five or ten times. This is quite handy. It's often the only way you can spread out a signal at the upper limit of your 'scope enough to really look at it. You'll pay a price in accuracy, but that's to be expected. Don't be surprised if the accuracy spec at the fastest sweep speeds isn't nearly as good as it is on the other ranges. That, also, is to be expected.

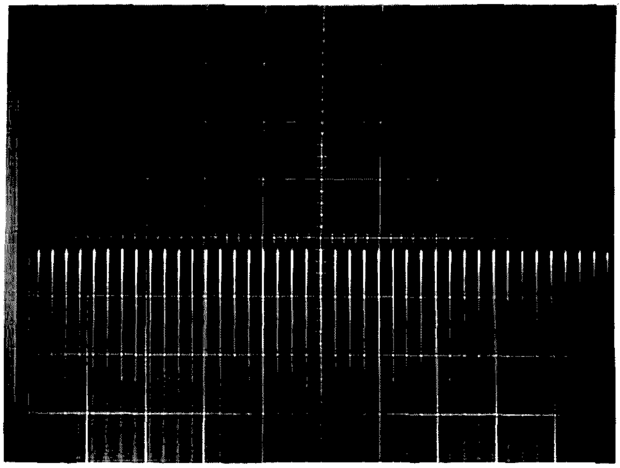
Delayed sweep is a commonly misunderstood feature. When I called on customers for a major instrument supplier, I was always amazed at the number of people who felt they positively, absolutely had to have delayed sweep — even though they didn't have the foggiest idea what it was or how to use it. If you really need it, nothing else will do. If you don't, save your money. Delayed sweep adds significantly to the cost of an oscilloscope.

Delayed sweep means that an instrument has two independent timebases. The second sweep runs after a delay of your choice, using the first timebase. The most useful implementation of this involves a precision, multiturn pot.

Why the precision pot? It lets you make a more precise time measurement than is possible on the screen. You don't have to worry about CRT nonlinearity, so you can get down into the 1/2-percent range by reading time off the pot rather than the screen. This can be important if, for example, you're trying to measure timing on a digital circuit. With the instruments available today, there are often better ways to make timing measurements without the expensive precision pot.

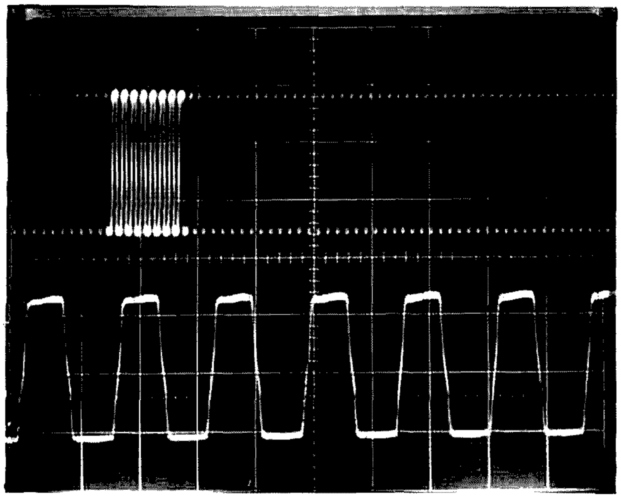
Another use for delayed sweep is found in digital work. Suppose you have some sort of index pulse, followed by a hundred or so data pulses. Let's also suppose that you want to look at each pulse in turn. A convenient way to do so is to use the delayed sweep. Let me illustrate.

PHOTO H



RF generator and frequency counter used to check time base accuracy.

PHOTO I



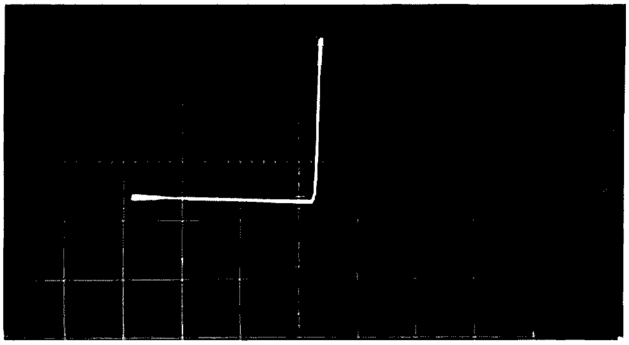
Upper trace: main sweep with an intensified zone (A intensified by B). Lower trace: delayed sweep alone. This photo required a double exposure. Some oscilloscopes are capable of actually presenting both traces on screen at the same time.

Photo I shows an upper trace with an intensified zone. The whole trace is a product of the main timebase. The intensified zone is a product of the delayed (second) timebase. The starting point of the intensified zone depends on the pot setting, regardless of whether it's a precision pot or not. The length of the intensified zone is the length of the delayed sweep. Changing from "main" or "A" timebase to "delayed" or "B" produces the result shown in the lower trace — an expansion of the intensified zone.

If you want to examine a pulse train, set it up as shown in the upper trace and look at the greatly expanded pulses using the B sweep. A good example of an application of this kind is viewing data off hard disks.

Once again, I'd advise you not to spend the money for delayed sweep if you don't have a use for it. The sweep magnifier lets you do almost the same thing on a more limited scale, and for some applications it's just as satisfactory.

PHOTO J



Current versus voltage curve of a diode, using the component tester in an oscilloscope.

Triggers

The most basic trigger circuit should have AC and DC coupled triggers, and a choice of positive slope or negative slope. This permits locking on the front or back side of most signals. These basic functions take care of a surprising number of situations. My own instrument spends practically all of its time in the AC+ position.

There are, however, some nice additions that don't necesssarily cost a lot. These are AC low frequency reject, AC high frequency reject, line, single sweep, TV horizontal, and TV vertical.

AC high frequency reject is useful if you need a stable display of an AM signal and want to look at the modulation envelope. With simple AC or DC coupling, your 'scope is going to have a hard time. Each RF cycle evokes a trigger, and you won't be able to stabilize the envelope. Switch in the high frequency reject, and the RF usually becomes a blur. But the envelope is stable, because that's all the trigger circuit can "see."

The reverse situation is also useful. If, for example, you want to look at an audio frequency signal that's carrying a bunch of RF trash, you can invoke AC high frequency reject and the 'scope will trigger on the AF only.

Line triggering locks you to the AC line. Want to know if a signal is derived from the AC line? Just switch to the line triggering, and look to see if your signal stands still on the screen. If it does, your signal is line related.

Single sweep doesn't rearm the trigger after the sweep. It lets you wait for a trigger, generate one sweep, and then waits for you to rearm the sweep manually. This gives you a convenient way to monitor for a single shot event.

TV horizontal and vertical triggers are just that. They let you generate a stable trace that's locked to the horizontal or vertical synch pulses of a television signal.

Extras or convenience features

The following features are ones that often make the difference between a tool that does the job and one that makes the tool a joy to use.

Consider a simple thing like graticule illumination. You'll notice that you can see the graticule in some of the photos, and in others you can't. The difference lies in the addition of a couple of small light bulbs and a pot on the front panel. This isn't a big deal from the expense standpoint, and is a very useful feature — even if you're not planning to take photos. Recently I finished up a little keyer for portable use, and had to trim the timing components for an exact 3:1 dash/dot ratio. I turned on my 'scope and put the timebase on one of its slower sweep speeds. My 'scope is a Heath SO-4251, which is not a storage model, and I had to turn down the lights to see the afterglow of the trace. When the lights are low, you can't see the graticule unless it's illuminated, so I was very glad to have that convenience.

A component tester is another feature that can be very handy. This is a simple gadget that allows you to put a low amplitude swept voltage on a component, even when it's in a circuit, and lets you see the component's current-versus-voltage curve. The component tester lets you make some intelligent estimates of component values and quality. You can see the trace of a good diode in **Photo J**. One word of warning. It's very templing to use a component tester across the emitter base junction of a transistor. If you're not careful, this can have a very insidious effect. It will degrade the beta of the transistor you're testing,[3] and create problems you didn't have before. Emitter base breakdown apparently punches little holes in the junction; that's something you don't need.

You might also want to check out the tube's accelerating potential. The higher it is, the brighter the tube will be. On a 20-MHz instrument 2.5 to 3 kV is enough. This is often the voltage manufacturers use, because it provides satisfactory performance at a reasonable price. As bandwidth increases, so does your need for accelerating potential, since your timebase spreads your beam electrons thinner and thinner as you increase sweep speed. Seven to 9 kV are nice in a 40 to 50-MHz instrument; very high bandwidth units can have potentials up to 25 kV.

Conclusion

Though this isn't a complete discussion of oscilloscopes, I hope you'll find it helpful as a purchasing guide. The information presented here should enable you to make an informed decision and have some sense of where the oscilloscope you choose can be properly applied. hr

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1. *The 1989 ARRL Handbook for the Radio Amateur*, ARRL, 1989, page 25-26.
2. Bob Orwiler, "Vertical Amplifier Circuits," *Tektronix*, 1969.
3. Wes Hayward and Doug DeMaw, *Solid State Design for the Radio Amateur*, ARRL, 1986, page 24.

More RIT/XIT for the Kenwood TS-530S

We all have our weak spots, and mine is CW DX. I don't rabidly chase DX, but when it's there I do my best to work it. So when I heard the XF4 expedition on 15 meters calling CQ and listening up 5 kHz, I found the frustration almost unbearable from not being able to tune that far from my transmit frequency.

The problem

The TS-530S is a serviceable rig, but it has a range of only about ±1.5 kHz on its receiver incremental tuning (RIT) function. Because RIT is usually implemented with a varactor and pot for bias, I thought I'd try to find it on the schematic in the owner's manual and see if I could coax it into tuning farther. Stripped of a lot of extra lines on the drawing, the schematic is shown in Figure 1. Pot VR5 on the IF assembly sets the range midpoint, and the varactor in the VFO assembly is the actual component that changes.

The "cleanest" modification would be to change the varactor for one with a wider tuning range. This would be hard to do, though, as the VFO is a sealed assembly. A cheap and dirty alternative would be to increase the voltage swing to the varactor, and this is the method I used.

Not knowing anything about the particular diode used, I was concerned about increasing the diode voltage. A quick look at the schematic showed that this wasn't possible, because the full 9-volt supply is across the diode when the pot is at one end of its rotation. All I had to do was increase the voltage swing.

If I knew the varactor tuning curve, calculating the range would be easy. Lacking that, I clipped some test leads across the 6.8-k resistor to ground and paralleled some resistance until the tuning range was what I wanted. I settled on a 1-k resistor as the value to add in parallel to the existing part.

Circuit changes

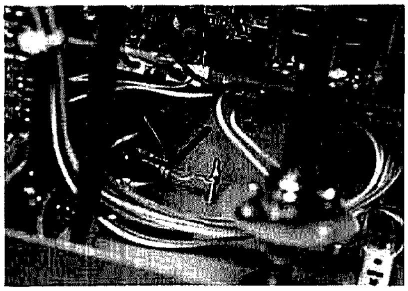
Installing the 1-k resistor (marked * in Figure 1) is the only thing tricky about this mod. First, I moved the existing resistor slightly out and down. Then I prepared the new part by forming a loop in one lead, tinning it lightly, and cutting the other lead to length. I looped the tinned end over the existing resistor, flowed the solder with my smallest iron, then wrapped the other end in place and soldered it. All of this inside work was made easier because I had a pair of hemostats* to do the fine positioning and holding. Photo A shows the completed modification.

Once the resistor is installed, you'll need to set VR5 so the frequency doesn't change when you turn on the RIT. Center the front panel pot and turn the RIT on and off. Note the frequencies with and without the RIT, and adjust VR5 so there's no change when you turn RIT on.

With the modification in place, the RIT tunes ±5.1 kHz. I was a little ner-

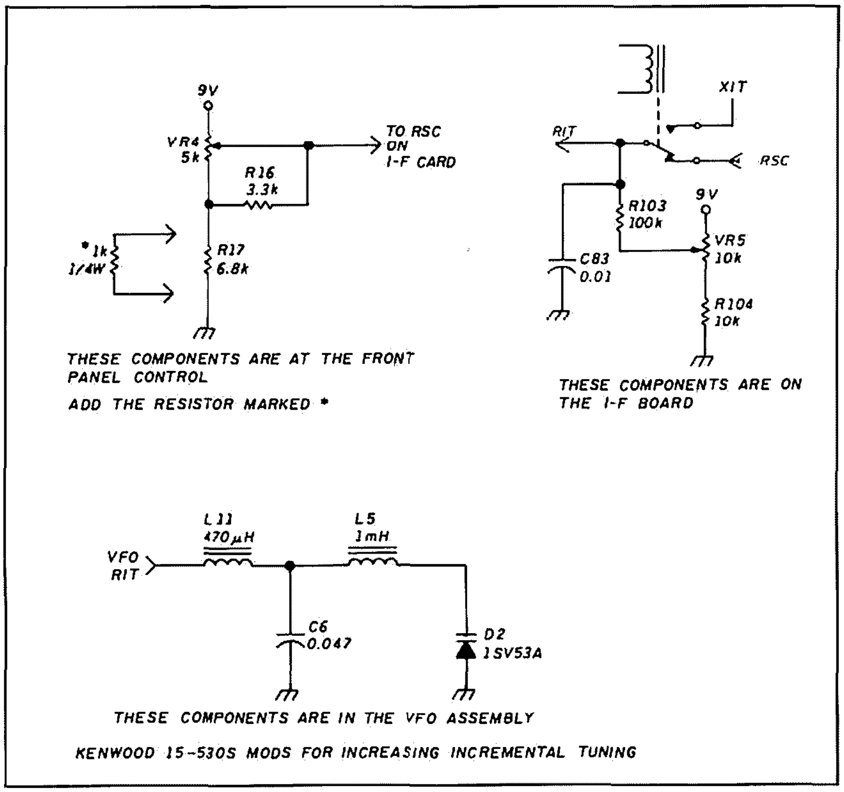
*A clamplike instrument used in surgery to reduce or prevent bleeding. Ed.

PHOTO A



Changes to extend the Kenwood TS-530S RIT take about an hour.

FIGURE 1



Kenwood TS-530S mods for increasing incremental tuning.

vous about trying for more range, because all but the most expensive varactors have a nonlinear tuning curve, and I wanted the resulting tuning range to be symmetrical on the RIT control.

This modification took about an hour from the time I decided to do it until the rig was buttoned up and operational. In case you were wondering, I worked the XF4 station a few hours later, both SSB and CW, with 5 kHz of RIT.

Bob Lombardi, WB4EHS

SKYWAVE COMMUNICATIONS PART 2

The atmosphere and the ionosphere

By Cornell Drentea, WB3JZO

According to Kenneth Davies,[1] Watson-Watt was the first to apply the term *ionosphere* to "that part of the atmosphere in which free ions exist in sufficient quantities to affect the propagation of radio waves." Other names associated with the ionosphere include Stuart (1878), Schuster (1889), Heaviside (1902), and Kennelly (1902). The discovery of how the ionosphere affects radio communications can also be attributed to the many Amateur Radio operators who have experimented with it during this century. The ionosphere can be a very hard thing to define due to its ever changing refractive and reflective characteristics. Although there is not complete agreement, it is generally accepted that the ionospheric region extends in the atmosphere from about 30 km (18.6 miles) above the earth's surface to almost 1,000 km (620 miles).

Assuming that its content remains constant with the altitude (a theory which is disputed), the *general atmosphere* is made up of the following gases:

Gas	Volume (percent)	Mass (percent)
Nitrogen	78.09	28.02
Oxygen	20.95	32.01
Argon	0.93	39.96
Carbon dioxide	0.33	44.02

Small amounts of water vapor, ozone, nitric oxide, and hydrogen also occur.

Within this concept, the atmosphere follows general meteorological behavior. Turbulent mixing exists up to 100 km (62 miles), with pressure and density variations behaving according to well-known barometric equations.

Some assume that above this altitude the atmosphere presents the same composition, but that its content is much more rarefied. Here, the ultraviolet rays and the x-rays coming from the sun generally produce the ionospheric reflective and refractive effects. During periods of disturbed sun, particle radiation — like high energy protons, alpha particles, and low energy protons and electrons — disturbs the ionosphere down to altitudes as low as 45 km (28 miles) above the surface. Under these conditions absorption of earth-generated waves takes place, so their energy will never reach the higher altitude reflective layers.

Sunspots and flares

The mechanism of the sun causes the *ionization of the* rarefied gases when the radiation passes through the ionosphere. This gaseous globe of fire is 540,000 km (335,540 miles) in diameter. Its volume is one and a third million times the volume of the earth, while its density is only one-fourth that of the earth. The temperature at the center of the sun is believed to be approximately 7.2 million degrees Celsius (13 million degrees Fahrenheit). There are those who dispute this figure, believing that it is twice that much. At the surface of the sun, the temperature has been measured in the vicinity of 5,500 degrees Celsius (10,000 degrees Fahrenheit). The sun's hot gases are composed of dissociated atoms. Elements like hydrogen make up about 70 percent of the sun's mass, while helium makes up about 28 percent. The remaining 2 percent consists of all the heavier elements. Only a few hardy compounds hold together in the unspotted areas of the sun. The light and heat radiation received continuously on earth from the sun is the equivalent of about 1.5 kW per square meter. The sun has been outputting this energy for more than a billion years, and will likely continue to do so for several billion more.

The most important process on the sun impacting the ionosphere is the formation of sunspots. Sunspots are confined, eruptive masses of cooled gaseous plasma, which exhibit molecular (versus atomic) composition and move about the sun's surface in a swirling fashion. This happens to a greater extent every eleven years or so, creating what is known as the sunspot cycle. However, a certain number of spots exist on the surface of the sun at all times. To an observer, they appear in groups and move in a westerly direction across the sun's surface. Their speeds differ slightly, depending on their exact latitude. The sunspot rotation cycle is approximately 27.5 earth days (the sun's rotation period), making it easy to predict a rounded 28-day

propagation cycle. Sunspot diameters have been measured at up to 128,000 km (79,535 miles). The spots appear darker than the surface ot the sun because of their lower temperature — 4,400 degrees Kelvin (4,126 degrees Celsius, or 7,460 Fahrenheit) compared with 6,100 degrees Kelvin (5,826 degrees Celsius, or 10,520 Fahrenheit). They are likely to appear in two rather narrow bands located equidistant from the sun's equator. Groups that are usually characterized by two larger spots accompanied by a number of smaller ones appear at latitudes of 5 to 30 degrees north and south of the sun's equator. Through selective spectrum analysis, the paired sunspots have been identified to contain reversed magnetic polarity from each other. For example, during a particular cycle the preceding spots in the Northern Hemisphere show their positive poles, while the following spots show their negative poles. The process is reversed in the Southern Hemisphere. An interesting phenomenon occurs when a new cycle begins; this magnetic mechanism appears to be reversed from the previous cycle.

Study of the molecular composition of sunspots has revealed that they contain about 18 compounds. These include titanium oxide and the hydrides (compounds of hydrogen with other elements) of calcium and magnesium. The sunspot's powerful magnetic fields have been measured at about 4500 gaus. They keep the colder matter confined within the spot's perimeters.

Explosive outbursts of matter and energy at the edges of the sunspots are called flares. These explosions are associated with sunspots, but not to be confused with them. Flares burst out in space from the edges of sunspots to an altitude of about 20,000 km (12,427 miles) at speeds of about 300 km per second (186 miles per second). The resulting radiation, composed of copious amounts of ultraviolet rays, x-rays, and cosmic particles, leaves the surface of the sun and travels out into space. Depending on its nature and, consequently, its speed, a certain amount of this energy arrives here on earth at various intervals. For instance, it takes approximately 8.3 minutes for the ultraviolet and x-rays to arrive in the ionosphere, 15 minutes for the high energy particles like alpha rays and protons, and 18 to 40 hours for the low energy protons and electrons. This latter type of radiation is also known as the solar wind.

Ionospheric interaction

As flare radiation penetrates the earth's atmosphere, it interacts with the rarefied gases starting at about 600 km (373 miles) from earth. The interaction with the denser gaseous layers located closer to the earth's surface tapers off the flare's energy at about 30 km (18.6 miles) from the earth's surface where the ionosphere ceases to exist. According to a recent theory, the ionosphere can be thought of as a thick layer of disassociated atmospheric gases impacted primarily by ultraviolet radiation and x-rays. Because the ultraviolet radiation is not monochromatic, you can assume that certain gases, depending upon how they occur at different altitudes in the atmosphere, provide various degrees of absorption to certain wavelengths. This, in turn, creates several layers of ionization.[2]

Another current theory says that, contrary to appearances, the ionosphere is not made up of separate physical layers, but that the degree of ionization (the number of electrons per unit volume) is a continuous function of altitude, with maxima and areas of inflection. Therefore, we should talk about "ionized regions" rather than the more accepted layer terminology.[3] However, the word layer would be acceptable (because it has been used for many years) if defined as the altitude at which wave reflection takes place.

In any case, the ionization process consists of a breakup of neutral molecules and atoms into free electrons and positive ions. The free electrons recombine with the positive ions or attach to other neutral molecules, creating negative polarized ions. It is believed that the free electrons are responsible for the ionospheric reflections and refractions. If you accept the second theory, electron density is a function of the continuous equilibrium between the electron creation processes and the de-ionization processes. It depends on the recombination of the positive ions with the electrons, and the electron capture by other neutral atoms. Because of their heavy mass, it is believed that ionized atoms do not play a role in propagation.

The status of the ionosphere is always changing and depends on many things like sunspot activity, the magnetic field generated by the earth, the time of day, and even the winds in the upper atmosphere. Proof of some of these dependencies was demonstrated during ionosphere heating experiments at Arecibo Observatory when it was discovered that plasma waves were being generated in the ionosphere. This in turn accelerated the free electrons to high energies, making them travel from one hemisphere to the other. The experiments at Arecibo have led to speculation that certain new forms of electromagnetic waves can be generated through the modulation of these electric currents into the ionosphere, and that the process can be used for communication purposes.

To conclude, the ionosphere is comprised of several layers of dense ionization which are always changing and overlapping each other, but that can be defined distinctly enough to be studied individually. Now I'll look at what happens to HF radio signals arriving in the ionosphere.

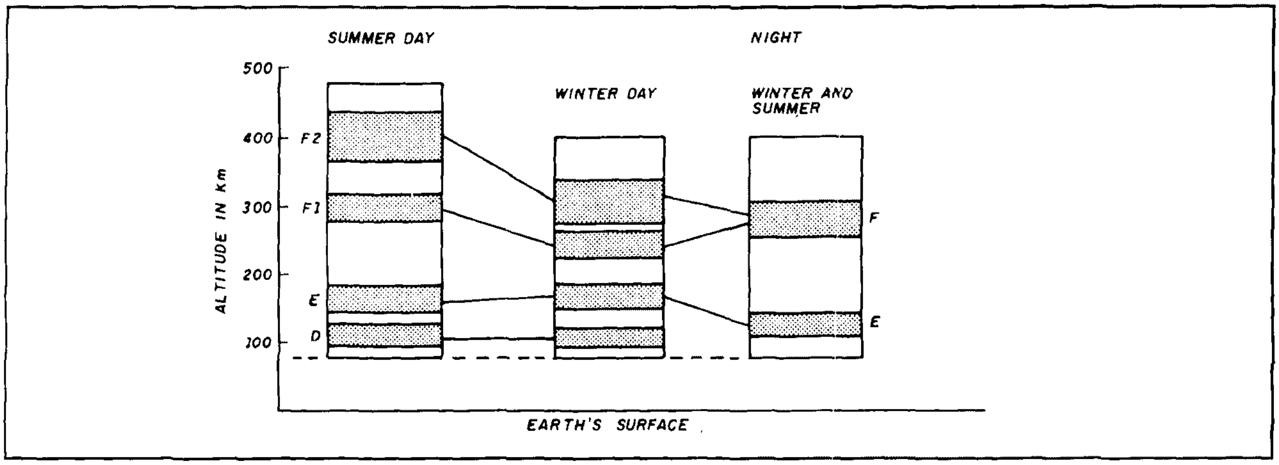
According to one theory, the electric field of an earth-generated HF wave arriving at an exited ionospheric layer causes the electrons to oscillate at the same frequency as the original signal. This oscillation produces a new electric field which is mathematically out of phase from the incident wave. The result is a perceived refraction or reflection of the wave at the ionosphere level. The transmitted signal can be received back on earth in the direction of the short or long paths at great distances. This is known as "skip." The regenerated (reflected) wave can bounce back and forth between the ionosphere and earth several times, making for audible and measurable delays over great distances (multiple hop propagation).

Some recent speculations suggest an even newer concept in long distance low power (QRP) communications. Some feel there is good reason to believe that the excited electrons in the ionosphere act not only as a reflector as is generally known, but as a plasma amplifier and a re-radiating antenna at the same time.* When conditions are right, a gain of anywhere from about 3 to 6 dB is believed to be obtained. The theory behind this mechanism is based on a lasing-like** reaction of the RF-excited mag-

* This theory was advanced by Albert Henderson, K6AJ, at RF Expo 1985 in Anaheim, California. It has not been validated by other laboratories, like the National Astronomy and Ionosphere Center, or by the Los Alamos National Laboratory.

** LASER stands for light amplification by stimulated emission of radiation; MASER stands for microwave amplification by stimulated emission of radiation.

FIGURE 1



Altitude changes for D, E, F1, and F2 layers depending on season and time of day.

netic fields, which in turn excites other electrons under the influence of the ultraviolet radiation in sort of a chain reaction.

The ionospheric layers

I've discussed how the ionospheric layers form. I'll describe these layers and their properties only briefly, assuming that from reading other publications you are generally familiar with the position and properties of each layer.

Sir Edward Appleton was responsible for defining and classifying the ionospheric layers in 1924. According to Appleton, the most important layers for communications are the D, E, F1, and F2 layers. The first layer identified was named E for the word electric. The other nomenclature followed in place. **Figure 1** shows how these layers change altitude and combine, depending on seasonal and time of day parameters.

The D layer and the lowest usable frequency (LUF)

The D layer is predominant between 55 to about 90 km (34 to 55 miles) and is a daytime event. At night the D layer disappears totally because the sun's radiation is stopped by the earth's shadow at these low altitudes. This layer is not very good at reflecting electromagnetic energy. Because gases are relatively dense at these altitudes, the ions produced in this layer encounter quick recombination processes (a lower number of free electrons) which produce energy losses rather than RF reflections and refractions.

In addition, the reflectance of this layer (if any) tends to be of a high angle because of its low altitude. Signal absorption occurs during the day at the lower HF frequencies. The frequency at which the D layer begins to allow HF waves to continue on to the higher reflective layers is known as the lowest usable frequency (LUF). This frequency varies with the sun's activity and can be shifted so high as to render a total blackout for all HF waves during periods of sunstorms.

E layer and sporadic E propagation

The E layer extends from about 90 to 125 km (55 to 77 miles). With its lower air density and higher angle of reflectance, the E layer acts as a better reflector or refractor than the D layer for relatively short paths. Because of its higher altitude, the E layer tends to produce results from some time before sunrise to after sunset, with minimums around midnight. Therefore, the E layer is primarily a daytime layer. Sporadic ionized patches of up to 150 km (93 miles) in diameter sometimes tend to form at the bottom of the E layer (about 100 km or 62 miles), providing powerful reflective and refractive properties for waves up into the very high frequency (VHF) range. It is believed that this phenomenon is caused by the interaction of gaseous winds in the atmosphere with the earth's magnetic field. It is called sporadic E because it only lasts for a few hours and occurs randomly. When present, sporadic E is responsible for spectacular communications at distances of up to 1,000 km (621 miles) for HF and especially VHF frequencies. Sporadic E clouds appear to drift with the upper winds, generally in a westerly direction and at speeds of about 150 km per hour (90 miles/hr). This type of propagation occurs to a greater extent at lower earth latitudes, commonly 20 to 40 degrees. Up to 100 hours of sporadic E propagation has been reported yearly in the United States during the months of May through August.

F layer, the maximum usable frequency (MUF), and the optimum frequency of transmission (FOT)

The F layer is used for long haul skywave communications. Its altitude is between 150 to 500 km (93 to 310 miles). Consquently, the F layer is almost always exposed to the sun's radiation. It is composed of two well-defined sublayers, F1 and F2. The F1 layer is the lower of the two, and extends from 150 to about 250 km (93 to 155 miles) during the day. The F1 layer behaves much like the E layer; that is, it achieves its maximum ionization at noon and disappears almost completely at night. Its altitude changes seasonally between 350 km (220 miles) during the day in the winter to about 500 km (310 miles) during the summer daylight hours. The nighttime altitudes are between 250 to 420 km (155 to 260 miles). Because of this, skip distance varies with seasons and time of day. Due to its height, the F2 layer is

always present but may be blocked to skywave communication by the absorption of the D and E layers. Efficient reflections and refractions can be obtained from the F2 layer year round.

The F2 layer is partly responsible for determining the optimum frequency of transmission. The art of predicting the FOT at any given time is dictated by complex calculations related to how the F2 layer impacts the maximum usable frequency (MUF) and how that relates to the lowest usable frequency (LUF) determined by the D layer. In practice, the FOT is taken as 85 percent of MUF. The MUF determination is also obtained from complex calculations based on standards set by the Comite Consultatif International des Radiocomunications (CCIR). Some of these standards, along with empirical data and exhaustive calculations, can be found in computer programs which readily predict propagation for the long term. They provide a means of planning radio communication systems and frequency schedules for months in advance while anticipating their variability under both quiet and disturbed conditions. Short term forecasting uses information derived from solar observations for day-to-day management of radio circuits. Long term forecasting takes into account the irregular behavior of the slow arriving particles known as the solar wind, and their variable interaction with the earth's magnetic field in the magnetosphere and the ionosphere. I'll discuss some of these programs in part 3 of this article.

The grayline and auroral propagation

One of the more interesting propagation modes is the *grayline. The grayline is a path created along a 500 km* (310 miles) wide area separating day from night, usually called the terminator. Enhanced north-south propagation along this path is possible depending on the earth's diurnal tilt (arc varies up to 47 degrees relative to the poles in one year). The HF spectrum affected by this type of propagation is generally from 2 to 10 MHz. Simple computer programs (which I will not discuss here), that anticipate how this tilted line crosses different areas on earth, can predict with some certainty when and where communication can be established. In grayline propagation, the D and E layers are actively providing strong absorption to the waves being transmitted on the day side of the line. On the nightside these layers are nonexistent, with a MUF which is usually below the frequency of interest. The energy is then channeled up toward the reflecting layers by these two conditions, in much the same way that light would be radiated upward from inside a narrow canyon. The energy is finally reflected back to earth by the F layer on the short or long paths along this grayline. Grayline propagation is useful for HF communications.

Another interesting propagation phenomenon is based on the aurora. During a magnetic storm, the solar wind particles, and particularly the electrons arriving in the ionosphere, become trapped by the earth's magnetic field in the polar regions. Here is how the entire process happens. As the earth spins, its crust moves faster than its core. This is a motor-like mechanism responsible for creating the magnetic field which surrounds the earth. As discussed earlier, the magnetic field around the earth is known as the magnetosphere. The magnetosphere with its well-known Van Allen belts is impacted in turn by the solar wind. The interaction of the solar wind (for example, electrons) with the magnetosphere turns into electricity. It has been estimated that more than a million megawatts are being generated at any given time, making gases glow in the dark or become fluorescent. If you believe the earlier theory that different gases concentrate at different altitudes, then electrons spiraling down the magnetic lines of the magnetosphere in the oxygen region will glow red or green. At the lower levels, the nitrogen molecules will produce violets and blues. Other gases, like hydrogen (in lesser amounts) will produce oranges, and yet a different kind of red. This is the aurora. HF signals are generally absorbed by the aurora. A strange fading occurs making the signals flutter drastically at a high rate (arctic flutter). It is possible to bounce signals off the aurora curtains with occasional success. A new kind of propagation known as the Auroral E has been used recently at VHF frequencies. This is a mode which is similar to sporadic E, and has been used at frequencies of up to 144 MHz.

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2. D.R. Dorsey, "Radio-wave Propagation — A Tutorial", Proceedings of RF Technology Expo 88, Anaheim, California.
3. Lucien Boithias, Radiowave Propagation, McGraw-Hill Book Company, 1987.

Ham Radio Bulletin Board

Dear Readers:

Welcome to the *Ham Radio* Bulletin Board. The "Bulletin Board" is our new forum for sharing some of the more technical correspondence we receive here at HR, additional information on articles we have run, or any other pertinent bits of news. This new feature will appear in the magazine whenever we have messages for you, or you have messages for us. You'll find the "Bulletin Board" interesting, thought provoking, and informative.

Terry Northup, KA1STC, Editor

Enhancing accuracy using insulating spacers

Hear HR:

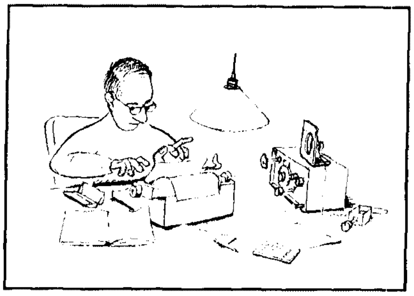
I really enjoyed the article in the January 1990 issue of *Ham Radio* on "Copper Pipe Transmission Lines." However, I would like to add a little information that will enhance the accuracy of the results.

The impedance values in **Table 1** or **Equation 1** are based only on air as the dielectric separating the inner and outer conductors. Many times it is necessary to use insulating spacers or beads in conjunction with air in order to maintain the concentric relationship between the conductors. When non-air or non-vacuum insulating material is introduced, this effectively raises the dielectric constant to some value greater than 1 (the value for air) and also effectively lowers the value of the characteristic impedance.

With this in mind, the following general rules should be used when insulating spacers are introduced with air dielectric:

1. Use the thinnest insulator spacer dimensions possible and the smallest number of spacers possible (the total spacer volume must be much less than the total air volume). Use low loss insulating material.
2. Always select a characteristic impedance value from **Table 1** that is slightly larger than the desired characteristic impedance.

Hilton A. Turner, Jr., KB8LS, Kokomo, Indiana

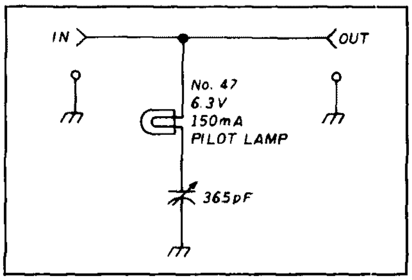


Being creative at little expense

I found a great and pleasant bit of nostalgia in Charlie Tiemeyer, W3RMD, and his article in the December 1989 issue of *Ham Radio*. The article is on page 42 and is titled "The Five Band Junkbox Transmitter."

Many times during my Ham career I have whipped up a "rigg" of this type, *but* with one addition!

Taking tips from articles in *QST* (Hyde, April 1955, page 51, and McCoy, September 1956, page 22) I always added a simple output indicator. The sketch I've included is self-explanatory (see below). I have built it into the transmitter or separately in a coffee can.



To use, simply set to minimum capacity and tune the transmitter normally. Next, increase the capacity *slowly* until the lamp barely lights. Then retune the transmitter for maximum brilliance of the lamp. This represents maximum output even though the meter may not agree.

Be my guest to add this device to your readers' fund of knowledge.

Do issue more articles of this nature. The beginners to our craft need to know that a fortune is not necessary to enjoy Ham Radio!

Gerald R. Skinner, K4LVZ, Winter Park, Florida

Theory and principle

Dear HR:

The article in the November issue ("A Low-Noise Design Primer," page 80) by Bob Lombardi, WB4EHS, on the subject of noise was very interesting but I would like to correct an historical reference.

Thermal noise in electrical circuits was discovered by J.B. Johnson in the course of research on telephone lines. The theoretical explanation of the noise (**Equation 1**, et sequence in Lombardi's article) was developed by H. Nyquist. Both Johnson and Nyquist published their work in the journal *Physical Review* in 1927 and 1928.

In physics it is common to refer to the noise, itself, as "Johnson noise" in honor of the discoverer. Nyquist's result is usually called the "Nyquist theorem" because he derived it from fundamental principles, specifically from the Second Law of Thermodynamics and the quantum theory of radiation.

It might interest your readers to know that the importance of Nyquist's result ranges far beyond the direct application of his theorem. His theory was generalized (in the early 1950s) to apply to such diverse phenomena as fluctuations in dielectrics and in magnetic materials, to scattering of electromagnetic radiation and neutron scattering. Even today, the Nyquist theorem and its generalizations are among the few exact theoretical connections between the science of equilibrium thermodynamics and electronic circuits (which never operate at thermal equilibrium).

Walter A. Simmons, Ph.D, AH6HU, Honolulu, Hawaii

Article update

Here's an update to the article "Keeping an Eye on Your Sideband PEP," by John Fielden, GW5NAH (SK), that appeared in our September 1989 issue. The original article refers to balancing out op amp offsets but doesn't suggest how it might be done. For those of you not familiar with such techniques, the following method is suggested.

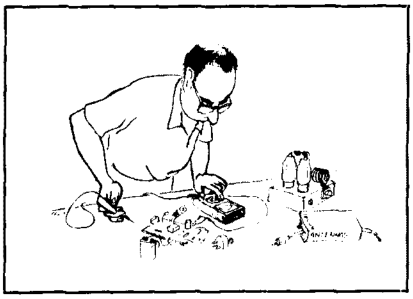
Apply a small DC voltage (about 2 to 4 mV) to the input of the module and

adjust RV3 to get the same voltage at the output. An accuracy of $\pm 100\ \mu V$ is quite sufficient. You can't make this adjustment with a zero input voltage as the output can't swing below about 1 mV. The settings of RV1 and RV2 don't matter for this adjustment, nor is the actual voltage used critical — as long as it doesn't exceed a few mV. Once the offset adjustment is completed, set RV1 and RV2 for the correct sensitivity as described in the article.

Probably the easiest way to get the small voltage needed is to connect the input of the module to the supply voltage via a resistor of 1 meg or so. You can use RV1 and RV2 to get a suitable voltage for the offset adjustment.

Bob Wilson, WA1TKH,
Consulting Technical Editor

Ham Radio Techniques



Bill Orr, W6SAI

THE YAGI OPTIMIZER

The year was 1946. There was plenty of exotic DX on the 20-meter band: AC4YN, XZ2KN, W6VKV/I6, ET1JJ, AC3SS, YI5AL, and lots more. The problem was to work it. And more often than not, I was beaten out in a DX pileup by my good friend and aggressive competitor Paul, W6MJB, a local DXer with a keen sense for the jugular.

"The reason he beats you out is that he has a four-element beam and you only have a three-element beam," observed W6VFR.

"The reason I beat you out is that I'm a better operator than you," said W6MJB with an engaging smile.

Very frustrating. Ol' Paul had put the hex on me for sure. Did he have a better antenna? Was he a more experienced DX chaser than I? Or both? I never resolved that problem.

But that was part of the fun. When, on a rare occasion, I did manage to snooker W6MJB, it really made me feel good. Maybe my antenna was as effective as his, after all!

Much has happened since those good old days, and a lot of the Black Magic once invoked in designing and building beam antennas has now become a science. Here's a case in point. A few weeks ago I was looking through a dusty old log book and found the dimensions and data for W6MJB's mighty four-element DX antenna. I entered the pertinent information in my modern computer and compared the results with similar data about the old three-element beam I used when chasing Paul around the 20-meter band. The results? Both beams performed about the same as far as gain and front-to-back ratio were concerned! There was no discernible difference in overall performance between the two antennas. Maybe Paul was right — he was a better operator than I was!

The Yagi Optimizer (YO)

Is a four-element beam a better performer than a three-element beam? What is the optimum boom length for a given number of elements? What is the effect of element taper? Is it better to lengthen the boom when adding an extra element, or to drop an extra element into an existing array in order to achieve a little more gain? What is the relationship between gain and front-to-back ratio? Is it necessary to sacrifice one to enhance the other? Do maximum gain and highest front-to-back ratio occur at the same frequency? These are good questions, ones that can finally be answered with the Yagi Optimizer program for the home computer.

The Yagi Optimizer Program* developed by Brian Beezley, K6STI, requires an IBM PC compatible computer with at least 512K of memory. You need a CGA, EGA, or HGC display. A math coprocessor is recommended, as is a hard disk, but neither is required. You should use DOS 3.00 or later to enable full access to all features of the YO program.

YO analyzes and optimizes a given Yagi antenna. It will model arrays of up to 50 elements. The model is examined in free space and the accuracy (compared with the MN program discussed last month) is typically within 0.1 dB for forward gain, a few dB for front-to-back, and within a couple of ohms for input impedance.

Optimization may be done at a spot frequency within a band, or at the low, middle, and high frequencies of the band of interest. You may choose the parameters, aiming for maximum gain, best front-to-back ratio, a given value of input impedance, or a combination of these parameters.

* Brian Beezley, K6STI, 507-1/2 Taylor, Vista, California 92084.

Is there a maximum gain Yagi?

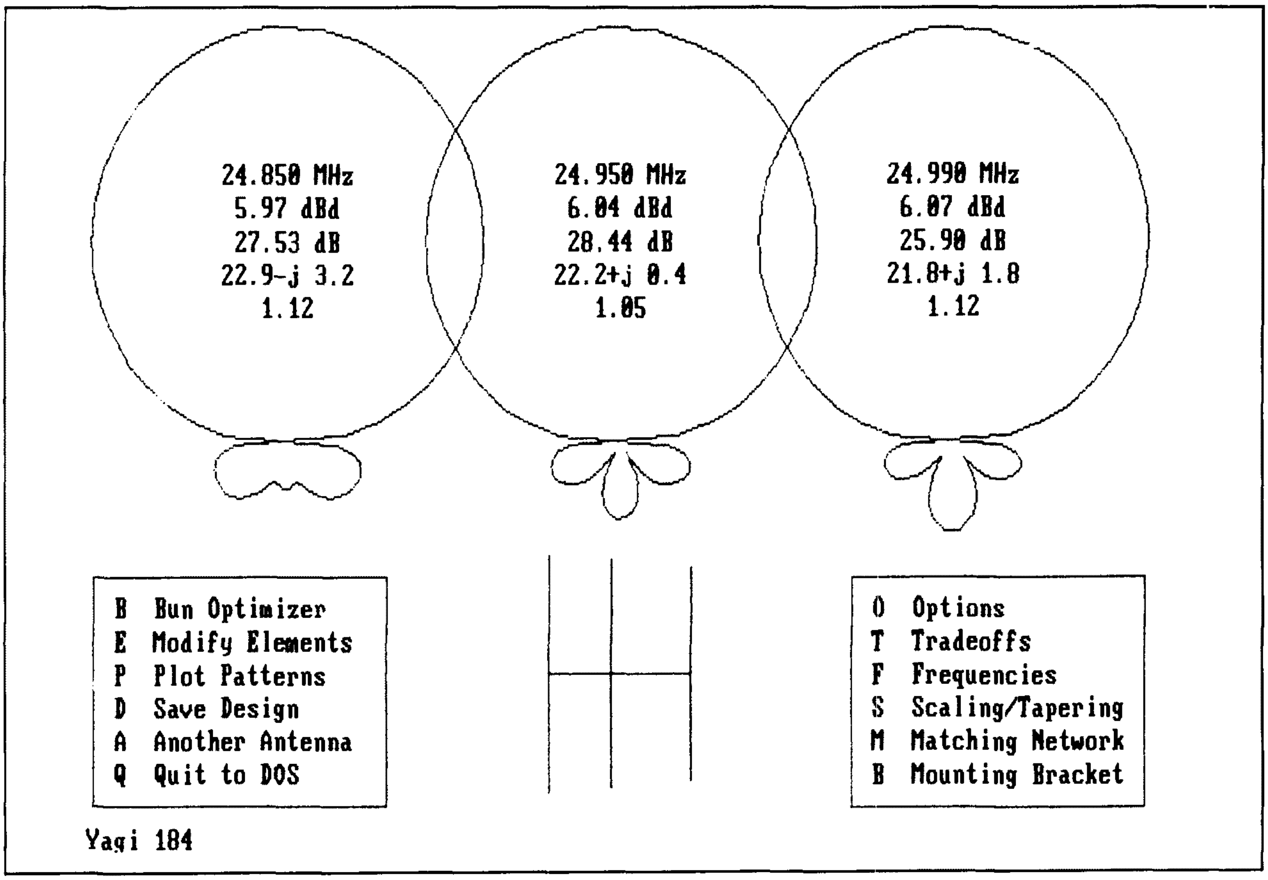
Before discussing program details, it's interesting to contemplate the possibility of achieving a maximum gain Yagi. A popular question among HF and VHF operators is: How much gain can be achieved for a given number of elements or a given boom length? YO will give interesting answers to such questions.

Yes, Virginia, there is a maximum gain Yagi. YO will find it for you by crunching through variations in element length and spacing that would be impossible to compute a decade ago. For example, I entered typical dimensions for a 24-MHz three-element Yagi built on a 12-foot boom into the computer. Starting with these basics, the program was directed to search for a maximum gain Yagi that retained a good front-to-back ratio while maintaining the same boom length. The results are shown in **Figure 1**. The program examined 184 length/spacing combinations to arrive at the optimized design. The array has 6.04-dBd gain at the design frequency of 24.95 MHz, and a front-to-back ratio of over 28 dB. Input impedance is about 22 ohms.

This is a good compromise design. Can the gain be increased by holding the boom length constant, but allowing element length and spacing to be varied? Yes. Setting YO in action again, I found that the program quickly examined 262 designs in sequence and arrived at the maximum gain design shown in **Figure 2**. This Yagi provides 7.82-dBd gain with a front-to-back ratio of only 7.55 dB. Input impedance is about 1.7 ohms. And look at the spacing!

A beam of this design is a theoretical concept. The radiation resistance is so low that substantial ohmic losses exist

FIGURE 1



24-MHz Yagi optimized for good combination of gain and front-to-back ratio. The figures in the patterns indicate frequency, gain, front-to-back ratio, input impedance, and SWR. Beam is optimized at three frequencies. Below the patterns you see a representation of the Yagi. Monotaper dimensions are (length and position in inches): reflector, 118.25/-72.96; driven element, 112.75/-9.23; director, 105.8/72.96. Boom length is 12 feet 2 inches.

in the array, and it's difficult to design a network that would match the low impedance load to a coax line. The real life gain of such an oddball design is questionable. Front-to-back ratio is poor and wind vibration would upset antenna parameters unless the driven element and reflector were suitably braced.

What's the upshot of this optimization? The design has gained 1.78 dB at the loss of nearly 21 dB of front-to-back ratio. At the same time, the feedpoint impedance has become impossibly low. Exotic results, but the beam is impractical. Optimization has been carried to an extreme.

Practical results with the YO program

Running the YO program quickly shows the folly of being too eager to achieve forward gain at the expense of other important antenna parameters. The YO program emphasizes that maximizing forward gain results in poor front-to-back ratio, low input impedance, possible sidelobes, and small SWR bandwidth. But because of the tradeoff capability, the program lets you automatically optimize a combination of gain, front-to-back, and either input impedance or SWR. The combination of parameters to be optimized and their relative importance are determined with a Tradeoff Menu.

The Tradeoff Menu

In most cases, maximizing forward gain at the expense of other parameters isn't practical. To obtain a good combination of parameters, set a ratio of forward gain to front-to-back of 9:1. This means that YO will weigh front-to-back by 10 percent and forward gain by 90 percent. (Other ratios may be chosen at your discretion.) This ratio, however, places 1 dB of gain equal to 9 dB of front-to-back ratio. If YO can change the design in such a way that the front-to-back ratio increases more than 9 times as much as the forward gain decreases, it will do so.

Experience has shown that this ratio yields a very practical design. Once this optimization has been completed, you can then optimize the design for input impedance and bandwidth tradeoffs. Your strategy is to obtain the most uniform antenna performance across the band. Obviously, the final design depends considerably on the width of the Amateur band; a 10-meter optimized design is quite different from a design optimized for the narrow 12-meter band.

The optimization method

The Yagi Optimizer uses a modified "Method of Steepest Descent." Readers with a calculus background

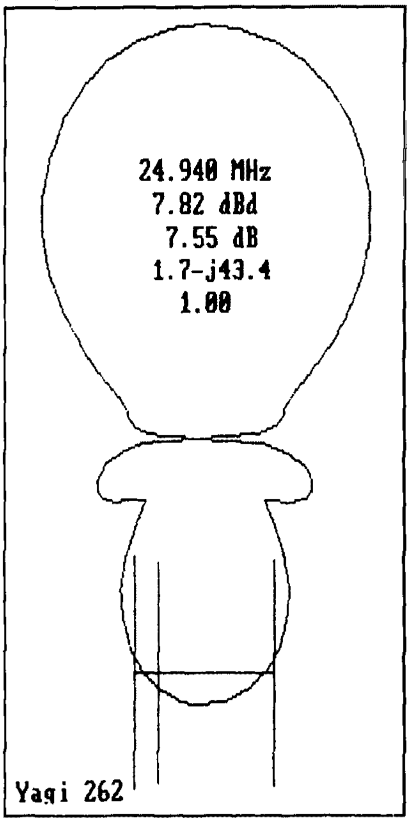
should recognize the technique immediately. Parameters of each element are changed by a small amount, while other antenna dimensions are held unchanged. The program calculates the sensitivity of the objective to each variable. The collection of the element sensitivities leads to an iteration where all the variables are updated, each in proportion to its respective sensitivity. The process is repetitive, and terminates either at user command or when no further improvement in the objective is possible.

This technique doesn't guarantee the very best set of antenna dimensions out of all possible dimensions. That nirvana can be accomplished only by an exhaustive and impractical search over all possible dimensions. The definition of "best" is user defined. As an analogy, imagine that you're climbing to the very top of a mountain *range of multiple* peaks, in a dense fog. You might arrive at a peak, but it might not be the highest one. Familiarization with the program and examination of various antennas in the program library will quickly guide you into a comfortable relationship with program tradeoffs. The program lets you change element lengths and positions by keyboard entry. This is a great way to get the "feel" for how each element affects the overall antenna performance. Common sense is a great help in Yagi optimization.

YO matching networks and other stuff

YO contains models for several matching systems including the Gamma Match, the T-match, the "Hairpin" Match, and the Beta Match. The program allows you to determine the effects of element taper and mounting brackets. It's also possible to quickly scale any Yagi design to a new frequency, while maintaining essentially the same performance characteristics. Finally, the program provides pattern plots which may be printed using a dot matrix printer with Epson compatible graphics. There are other interesting "bells and whistles" in this program, but I won't cover them here.

FIGURE 2



Same Yagi as in *Figure 1* optimized for maximum gain without regard to front-to-back ratio. Beam is optimized at only one frequency. Monotaper dimensions are (length and positon in inches): reflector, 115.251/0; driven element, 108.008/23.663; director, 111.006/142. Boom length is 11 feet 10 inches.

The YO file format

The initial antenna configuration must be entered in a Yagi file in a specific format. The elements are specified in order, beginning with the reflector (see **Table 1**). One-half of each element, from the center of the boom to the element tip, is normally required. For monotaper elements, you may use the whole element length.

You place a title for the antenna on the first line. One or three analysis frequencies go on the second line. The program assumes you are using megahertz if you don't specify frequency units. On the third line enter the number of elements and the dimension units. In this example, inches are used.

The next line asks that you list the taper diameters, starting with those closest to the boom. The maximum number of taper sections you can use is seven. The first taper section in this example represents the element mounting plates.

On the remaining lines, list the element positions followed by the length of the taper sections, beginning with the section closest to the boom. (As an alternative, you may use element spacings instead.)

It's possible to accommodate different taper schedules for different elements. Various examples are given in the documentaion that accompanies the program.

It's important to note that if you start out with a poor Yagi and optimize it, you'll get an optimized, poor Yagi. A modicum of common sense is vital in running a program of this type.

Used together, the MN and YO programs let you design and fine tune a Yagi array. You'll end up with a design that can be built, placed in final position, and used with a minimum amount of top-of-the-tower adjustment. I like that. As I've said before, I'm a coward when my feet are off the ground.

A final word

Here are some notes of caution. These antenna analysis programs and other programs of this type require that you have a good working knowledge of MS-DOS or PC-DOS. The programs aren't tutorials for running the complex data operating systems used by the IBM-type computers. Most of my early cockpit problems with antenna programs were due to poor insight into DOS management. Once I overcame

TABLE 1

YO format for 4-element beam.

K7HYR's maximum gain Yagi
24.890 24.940 24.990 MHz
4 elements, inches

	1.617	1.250	1.125	0.875
0.000	2.938	15.062	66.000	33.305
124.000	2.938	15.062	66.000	28.248
248.000	2.938	15.062	66.000	26.815
372.000	2.938	15.062	66.000	28.313

this hurdle, it was smooth sailing from then on. I'd also like to note that IBM clones are equal, but some are more equal than others. You may find occasional pitfalls when using a particular clone with an idiosyncracy that doesn't quite match the IBM machine the author used to prepare the program. Luckily, these "glitches" can be worked out as time goes by. In my case, a suggested print command on a different antenna program refused to provide a readable copy. A friend who is a knowledgeable computer programmer provided me with the solution in a few moments.

I believe that antenna modeling is a powerful tool for Amateur Radio. From time to time I hope this column can offer antenna designs that have been refined by computer analysis for the best possible operating characteristics. Antenna analysis is here to stay, but don't throw away that SWR meter just yet! hr

THE SUPER-DISCHARGER

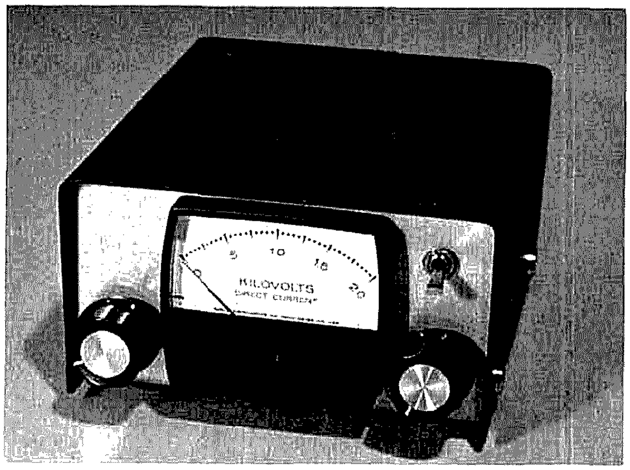
A simpler, easier to build test load for your NiCds

By W.C. Cloninger, Jr., K3OF, 4409 Buckthorn Court, Rockville, Maryland 20853

In a previous article[1] I discussed testing NiCds (nickel cadmium batteries) to determine their useful capacity. The process consists of charging the NiCds fully and then discharging them into a load at the mAh-times-1 rate. The multipurpose resistive load described[1] works well but requires close attention and several adjustments to maintain a reasonably constant load. Several full charge, full discharge cycles (1.0 volt per cell nominal minimum discharge voltage) can substantially improve the capacity of many NiCds.

During many NiCd discharge cycles, I've often wished that I had a constant current load that didn't vary with decreasing voltage. As a first attempt at a constant current load, I short circuited an existing current-limited power supply that uses a 723 regulator. It showed promise, but the input voltage versus load wasn't nearly as flat as I had hoped. This was probably due to the IC and the number of associated components. Then I discovered that it was best to keep things uncomplicated.

PHOTO A



The Superdischarger. The meter really indicates 0 to 2 A (but the price was right!). The two knobs are for coarse and fine-load adjustment.

Enter the Superdischarger

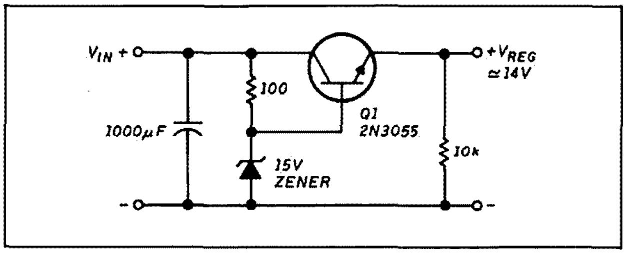
The circuit is simple, infinitely variable within its design range, smaller than the resistive load, doesn't require the special large resistor used in my earlier project,[1] and has a reasonably flat input voltage versus current load. The Superdischarger is shown in Photo A. It was specifically designed to discharge AA NiCds at 500 mA and sub-C, C, and D cells at 1.2 A (1200 mA). The load range of the Superdischarger extends from about 480 mA to just over 1.5 A.

Theory of operation

The Superdischarger is a current-limited power supply with an output load that will cause current limiting. A dead short of the output is the easiest "load" I could think of.

First, here's a quick lesson in one method of power supply regulation.[2] Figure 1 shows a basic zener-regulated

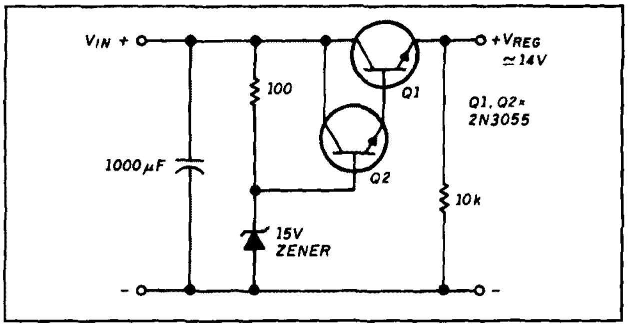
FIGURE 1



A basic zener-regulated power-supply circuit. It is not good for current loads over about 2 A.

power supply circuit using a single pass transistor. The maximum current output of this circuit is a couple of amperes, due to component losses and the gain factor of Q1. **Figure 2** is the same basic design; however, it uses a Darlington pair for the pass transistor. This circuit is good for maybe 5 A, but neither of these circuits has current limiting. I have seen both of these circuits used in commercial power supplies, usually less expensive ones.

FIGURE 2



An improved zener-regulated circuit using a Darlington pair, which improves usable current output to about 5 A.

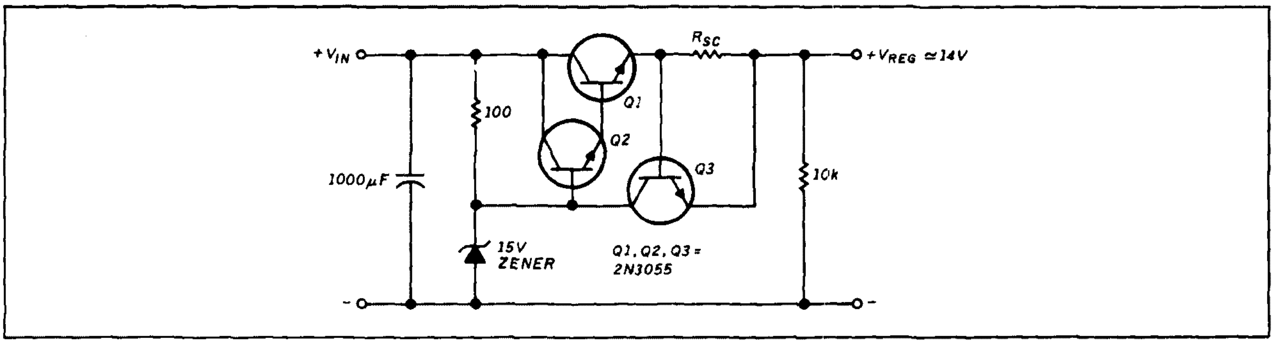
Figure 3 shows how current limiting is added to either power supply circuit. When the voltage drop across Rsc reaches about 0.65 volt, Q3 begins to conduct and removes current from the base of Q2, reducing the output to a fixed level. This is the method of current limiting used in the Superdischarger.

These circuits have been published many times and are suitable for some voltage regulation applications. Better power supply circuits may use the LM723 voltage regulator[3] (Astron, Drake P7, and others). Don't forget the "three-legged" fixed or variable regulators if they suit your needs.

Back to the Superdischarger

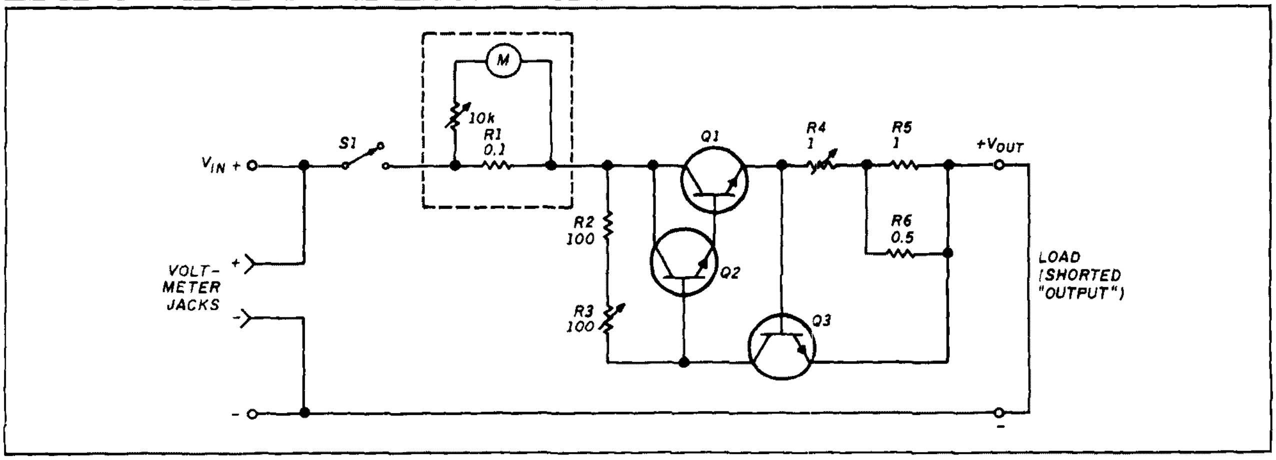
Because I needed only current limiting, I didn't use the zener diode. I now have an unregulated but current-limited "power supply" (see **Figure 4**). As the input voltage will vary with the voltage of the NiCd battery pack tested, voltage regulation isn't desired. My final design uses the Darlington pair because it will provide a 500-mA load down to 2.0 volts (2 AA cells). It won't maintain a 1.2-A load at 2.0 volts. You can use a single pass transistor, but the minimum voltage for a useful load is closer to 4 volts. There is one possible "advantage" to the single pass transistor version. You can

FIGURE 3



Current limiting is added to the power supply circuit with Rsc and Q3. All transistors are 2N3055, but only Q1 requires a heat sink.

FIGURE 4



The Superdischarger circuit will provide a 500-mA to 1.5-A load over a range of 3 to 15 volts. Potentiometer R3 is optional, but permits fine load adjustment. A 0 to 2 A meter would eliminate the meter network (0.1-ohm and 10-k resistors) and the 200-µA meter. S1 turns the load on and off.

even short circuit Rsc and the whole circuit will draw only about 2.5 A at 14 volts. Remember what I said about Figure 1 having a limited regulating capacity. I'm not sure that a single pass transistor version could ever self-destruct if a reasonable amount of heat sinking is used. The power transistors used are so cheap, why not use the Darlington pair?

Construction

Drill the project box and cut holes for the components. Pop rivet a terminal strip to the bottom of the box. You'll use this to support several of the components.

Mount pass transistor Q1 directly on the rear of the project box. I found that no additional heat sink was required (see **Photo B**). The Superdischarger does get warm, but not hot, for a 1.2-A load at 13.8 volts (about 16 watts of power dissipation). Solder Q2 directly to Q1; no heat sinking is required. Now solder Q3 to the terminal strip along with a couple of resistors and connecting wires as shown in Photo C.

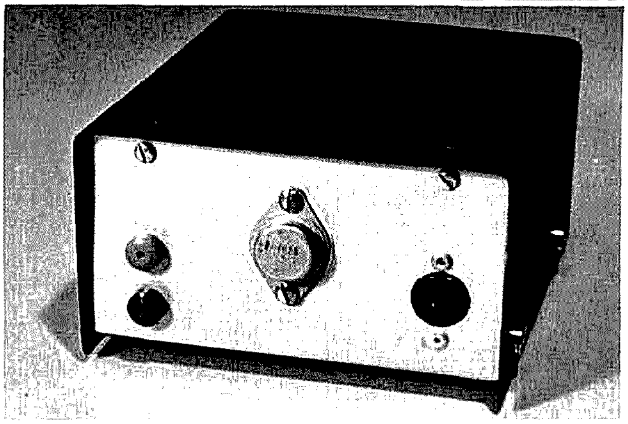
The meter was a hamfest special. It senses the drop across R1 to give the proper current indication. R2 is 2 watts or larger; R3 and R4 are 5-watt wirewound potentiometers. They cost about $10 each, new. My meter was large, so I had limited room to mount the potentiometers on the front panel, and used the miniature Mallory VW series pots. I suggest using a 2 or 3-watt wirewound pot. Normal sized pots are about $5. I recently found several at a hamfest for $1.50 each. R3 is optional and provides the fine adjustment of the load. R4 provides all the load adjustment you really need, but I'm just a fanatic!

Provide input connectors to suit your needs. I added tip jacks to the back panel. I plug the leads from my voltmeter into these jacks to check the voltage during the discharge process. The project box from Radio Shack has a nice steel top, but the aluminum chassis is quite thin. The Cinch-Jones connector I chose requires considerable effort to use, so I pop riveted a piece of aluminum angle stock to the top and attached the rear panel to it with sheet metal screws (see Figure 5). The top now acts as part of the heat sink.

Use and operation

Connect your NiCd battery to the Superdischarger, monitor the voltage, switch on the load, and adjust R4 to

PHOTO B

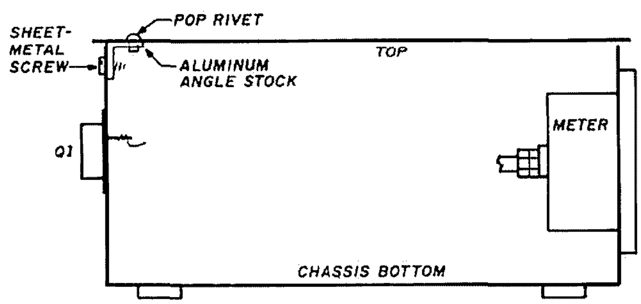


Back panel showing input jack and meter jacks. Q1 is mounted directly to the chassis (insulated if you wish) and requires no additional heat sink.

PARTS LIST

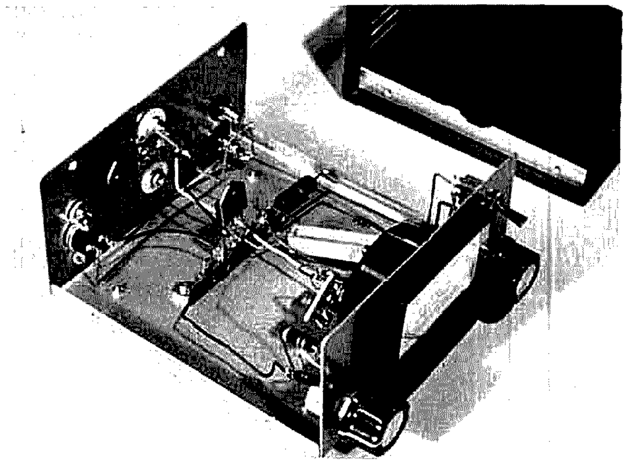
	Radio Shack part no.	Price
S1—SPST switch		
R1—0.1 ohm (optional depending on meter)		
R2—100 ohm, 10 watt	271-135	$ 2/0.99
R3—100 ohm wirewound pot (optional)		
Mallory VW-100 (5 watts)		about 10.00
Clarostat 58C1 (100 ohms, 3 watts)		about 5.00
R4—1 ohm wirewound pot		
Mallory VW-1 (5 watts)		about 10.00
Clarostat 58C1 (1 ohm, 3 watts)		about 5.00
R5—1 ohm, 1 watt	271-131	2/0.99
R6—0.5 ohm, 1 watt	271-139 (0.47 ohm)	0.49
Q1—2N3055 NPN, TO-3 case	276-2041	1.99
Q2, Q3—TIP3055 NPN, TO-220 case	276-2020	1.59
Metal cabinet	270-253	5.69
Terminal strip	274-688	4/0.89
1-inch knobs	274-416	4/1.79
Meter and hardware to suit your needs		

FIGURE 5



Sectioned view of the project box shows the aluminum angle stock and sheet-metal screws used to secure the top to the back panel for added strength and heat sink.

PHOTO C



Simple point-to-point construction is used. Q2 is mounted directly to the terminals of Q1. Q3 is mounted on the terminal strip.

500 mA or 1.2 A as needed. Make slight adjustments to keep the load constant as the battery voltage drops. Calculate the capacity of your battery by multiplying the time (hours) by discharge rate. If your AA cells give 500 mA for 1 hour, their capacity is 500 mAh or probably 100 percent of what they're rated (I won't discuss the 1-hour versus 10-hour ratings here).

I find that I'm easily distracted during the discharge cycle and tend to forget about the battery, so I use a kitchen timer and set it to go off every 5 minutes. I can then adjust the load and log the voltage and time, so I can plot a discharge curve if I wish. I also check and log individual cell voltages if I'm not discharging a sealed NiCd pack.

Now if I had a Superduper Discharger, which would automatically turn off the load at a preset voltage, log the time, graph the results... hr

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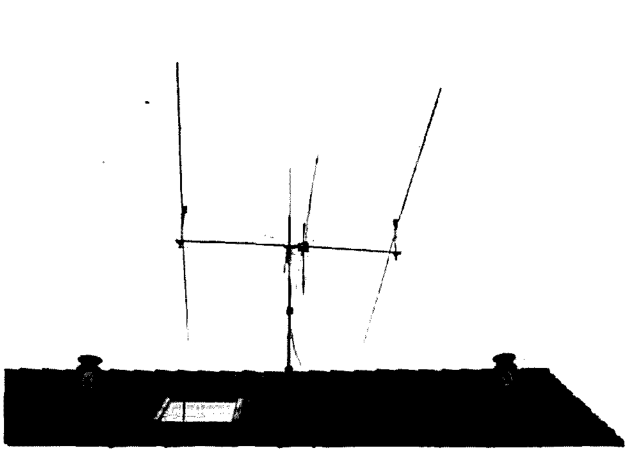
1. W.C. Cloninger, Jr., K3OF, "Get The Most From Your NiCds," *Ham Radio*, December 1988, page 88.
2. Paul J. Dujmich, WA3TLD, "Power Supply Regulation — Using Common Sense," *73*, January 1978, page 140.
3. W.C. Cloninger, Jr., K3OF, "723 Voltage Regulators," *Ham Radio*, March 1989, page 42.

OUTSTANDING 144-MHZ EME ACHIEVEMENT

Over 100 initial EME contacts with a two-Yagi station

By Bill Takacs, KB8ZW, 16724 Snyder Road, Chagrin Falls, Ohio 44022

PHOTO A



Twin 16-element Yagis perform very well for PA0JMV's EME experiments.

On a recent trip to Holland I had the pleasure of visiting PA0JMV's remarkable 2-meter, two-Yagi earth-moon-earth (EME) station. What Joe Mutter has done with two Yagis is almost unbelievable. Looking over his QSL cards and logs made me a believer. Joe has managed over 100 initial EME contacts with his two-Yagi station. He has worked 67 countries, 28 states, 55 four-Yagi stations and 2 two-Yagi stations. Photos A through D show this remarkable EME station.

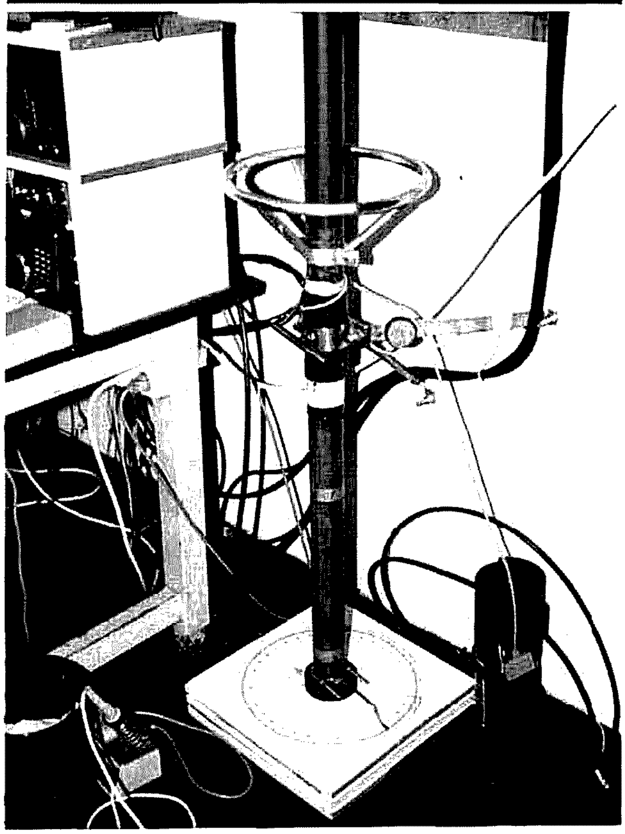
Antennas

Joe has kept to basics, and the numbers on VK3UM's EME program will demonstrate why this is possible. His antennas are modified homemade 16-element arrays, patterned after KLM's, stacked horizontally at the peak of his roof. Nineteen and one-half feet of coax from each antenna goes to a quarter-wave power divider and to a preamp coax relay. Another 20 feet of coax goes from relay to hamshack, giving 0.7-dB loss into the transmit coax. The preamp is part of a coax relay, with the connector on the relay incorporated as the input circuit of the preamp. He also uses a 70-dB isolation relay. The preamp measures 0.2 noise figure with a 3SK129.

Azimuth and elevation control

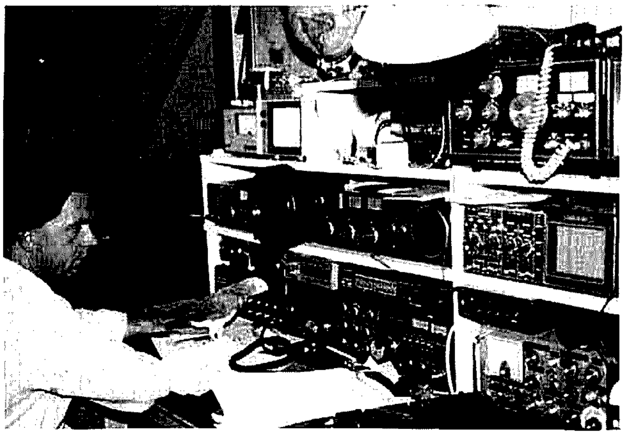
As demonstrated by his azimuth and elevation system, Joe believes in keeping things simple. Azimuth is obtained by turning a big steering wheel which rotates a mast that extends to the roof. The brake is a simple wing clamp affair that tightens the mast. A 12-inch, 360-degree compass rose on the floor gives azimuth readings. I questioned the elevation and Joe said, "Wait, let me get my elevation shoes." Puzzled at this comment, I waited while Joe disappeared downstairs. When he came back up with his elevation

PHOTO B



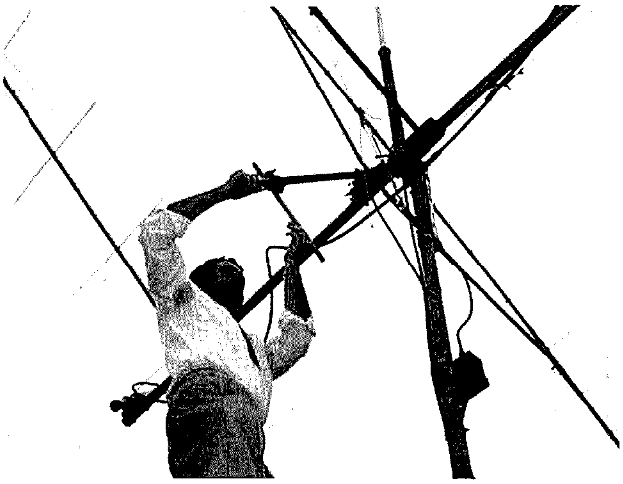
Antenna azimuth control.

PHOTO C



EME enthusiast Joe Mutter, PA0UMV.

PHOTO D



PA0JMV adjusts antenna for elevation control.

shoes, he demonstrated his system. To adjust the elevation, Joe opens the skylight, climbs onto the tile roof (hence the shoes), and tightens or loosens a tent guy line arrangement until proper elevation is achieved. The system is simple but effective. (I'd sure hate to do this at night during one of our northern Ohio winters.)

Electronics

Other equipment consists of a homebrew transverter with a GaAsFET 3SK124 to a Schottky mixer to 28 MHz. The power amp is made up of two 4CX250Rs driven by a BLY88. The power after a low pass filter is just over the kilowatt level. There are two HF transceivers; one is a JRC JST135 and the other a Yaesu FT102 plus FV102DM. Audio is processed through a homemade audio filter that is adjustable to 20 Hz.

Background data

Joe's interest in EME was sparked in 1973 when the United States Naval Research Lab (W3KE) started experiments with their big dish. He called them and was heard 459 plus sideband using a pair of F9FTs. They didn't transmit, so he made no two-way contacts. The day after, Joe's father looked at what he had built and asked him to "remove those dangerous looking things." In 1975, Stanford Research Institute organized an EME test weekend. This station was equipped for two-way contacts, but Joe decided that, with its 150-foot dish, it wasn't really an Amateur station. He had called W6PO the day before and asked him for a try. This was Joe's first 2-meter EME QSO. The next day he worked WA6LET without any problems.

Shortly after this, Joe's radio activity went to low-key operation. But in 1986, after some encouragement from SM2CEW, Joe again became interested in EME. A contact with W5UN using a ten-element CueDee and a kilowatt infected Joe with the EME bug, and brought him to where he is today. hr

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WANTED: Ham equipment and other property. The Radio Club of Junior High School 22 NYC, Inc. is a nonprofit organization, granted 501(C) (3) status by the IRS, incorporated with the goal of using the theme of Ham Radio to further and enhance the education of young people nationwide. Your property donation or financial support would be greatly appreciated and acknowledged with a receipt for your tax deductible contribution. Meet WB2JKJ and the "22 Crew" at Dayton, we will have two tables for "Classroom Net" members to congregate at. Talk in on 144.30 throughout the Hamvention. Join the "Classroom Net", 7.238 and 21.395 daily. Write us at: PO Box 1052, New York, NY 10002. Round the clock hotline: (516) 674-4072.

FOR SALE: ICOM 2400, ICOM IC-901. (201) 991-9280 after 7 PM EST. Also R-7000.

FREE CATALOG. Interfaces for IBM compatibles. Digital I/O and Analog input. Control relays, motors, lights, measure temperature, voltage. John Bell Engineering, Inc, 400 Oxford Way, Belmont, CA 94002. (415) 592-8411.

WANTED: Manual or copy for Simpson V.T.V.M. Model 266, Gonset Communicator VFO (6,2,220). Buy or swap. Alex Hellman, W2OEQ, 83-55 Woodhaven Blvd, Woodhaven, NY 11421.

DRAKE TR4 TRANSCEIVER, power supply and speaker, excellent condition. $325. Mosley CL33 tribander, $150. Rich Assarabowski, K1CC, 306 Vernon Avenue, Vernon, CT 06066. (203) 875-0166.

HAM SOFTWARE FOR IBM. Beam heading, grid squares and more. $5.95 includes shipping and free catalog. Personal checks accepted. Unicorn Software, 167 Kuntz Road, Hanover, Indiana 47243.

BATTERY PACK REBUILDING: Don't pitch it—mail it—for FAST—PROFESSIONAL service! Satisfaction guaranteed! ICOM: BP2/BP3 $19.95, BP5 $25.95, BP7/BP8/BP70 $32.95. KENWOOD: PB21 $15.95, PB21H $21.95, PB25/25H/26 $24.95. YAESU: FNB9 $19.95, FNB10 $23..95, FNB4/4A $36.95, TEN-TEC: $24.95 "U-DO-IT INSERTS" ICOM: BP3 $16.95, BP5 $22.95, BP7/8 $27.50, KENWOOD: PB21 $12.95, PB24/25/26 $19.95, TEMPO: S1-15/SERIES $22.95, YAESU: FNB4/4A $32.95, FNB10 $18.95, AZDEN: 300 $19.95, "NEW PACKS" ICOM: BP5 $43.95, BP8S $58.95, YAESU: FNB2 $19.95, FNB10S/FNB12 $44.95, SANTEC: 142/1200 $22.95. TELEPHONE/PAGER/COMPUTER/COMMERCIAL PACKS—Free catalog. Add $3.00 shipping/order. PA + 6%. VISA/MC add $2.00. CUNARD, RD 6, Box 104, Bedford, PA 15522. (814) 623-7000.

VACUUM VARIABLE- Jennings type UCS 10-300pf 7.5KV, new, $85. 813, $33.50; 4-250, $62.50. ARTTZ VE6GB, Box 3222, Airdrie, Alberta T4B 2B6.

"SOFTSTART". Protect your valuable tubes and poer supply diodes from DANGEROUS inrush surge currents. Can be retrofitted into most amplifiers. Completely assembled Amateur net for "SOFTSTART" is $49.95 plus $5.00 shipping in the 48 states. OMEGA ELECTRONICS, 4209 Live Oak Road, Raleigh, NC 27604. (919) 231-7373.

FREE LIST of low cost ham equipment. Write to Jim Braddy, WA4DSO, 3037 Audrey Drive, Gastonia, NC 28054.

WANTED: BUY & SELL. All types of Electron Tubes. Call toll free 1-800-421-9397 or 1-612-429-9397. C & N Electronics, Harold Bramstedt, 6104 Egg Lake Road, Hugo, MN 55038.

UPGRADING AND REPAIRING PC'S. A comprehensive resource to personal computer upgrades, repair, maintenance and troubleshooting. Covers all types of IBM computers and compatibles. 750 pages. $27.95. Personal checks accepted. Unicorn Software, 167 Kuntz Road, Hanover, Indiana 47243.

RADIO SHACK Color Computer Ham Software and Hardware. Free catalog. Dynamic Electronics, POB 896, Hartselle, AL 35640. (205) 773-2758.

HAM SOFTWARE IBM/Compatibles 10 disks $26.95. MC/VISA/Discover. N5ABV EAPCO/H, Bx 14, Keller, TX 76248-0014. (817) 498-4242. 1-800-869-7208.

WANTED: Original 40's and 50's service test equipment manuals. No HP or Tektronix. R. Galloway, 2220 Woodbury Rd, Melbourne, FL 32935.

WANTED: B&W 40TVH and 80JCL plugin coils. James Schliestett, W4IMQ, PO Box 93, Cedartown, GA 30125. (404) 748-5968.

VIBROPLEX COLLECTOR'S GUIDE. Covers all bugs and paddles. Includes history of each key, identification, dating tips, complete patents, and more. $14.95 plus $2.00 s/h (foreign, $5.00 s/h). Detains SASE. Artifax Books, Box 88-HR, Maynard, MA 01754.

YAGI BUILDERS. Tube traps, tubing, clamps and universal plate for boom to mast or elements. Six band rotary dipole or Yagi. 10thru 40M. SASE for details. 1-800-833-3794, 1-2 PM EST. Brown Engineering, Inc, 5501 SW 25th Court, Hollywood, FL 33023.

WANTED: Schematic/service manual for my Hallicrafter SX117-HT44 Hunter Bandit Linear. Bill Webb, K4OPE, POB 5, Paint Rock, AL 35764.

PRE-1942 TRANSMITTER WANTED. Any condition or part considered. Examples: Gross, Stancor, Collins, Thordarson, Harvey, Temco, National, Utah, Hallicrafters, Hammarlund, Guthman. Bob Mattson, KC2LK, 10 Janewood, Highland, NY 12528. (914) 691-6247.

CHASSIS AND CABINET KITS. SASE. K3IWK, 5220 Harmony Grove Rd, Dover, PA 17315.

HALLICRAFTERS SR400A CYCLONE III. 80-100 meters, 550 watts SSB, AC PS with speaker. DC PS. Mobile mounting bracket, manual. Shure 444 mic. $400. B.A. Fred, KA9OEK, 25W632 Flint Creek Rd, Wheaton, IL 60187. (708) 668-7460.

RTTY JOURNAL published 10 times per year for those interested in digital communications. Read about RTTY, AMTOR, MSO'S, PACKET, RTTY DX and Contesting. Plus technical articles concerning the digital modes. $12.50 per year (foreign higher). RTTY JOURNAL, 9085 La Casita Ave, Fountain Valley, CA 92708.

CUSTOM MADE EMBROIDERED PATCHES. Your design, any shape and colors. Free ordering information, price list and random sample. HEIN, 79602 Manitou Trail, Glen Arbor, MI 49636-9711.

IBM-PC RTTY/CW. New CompRtty II is the complete RTTY/CW program for IBM-PC's and compatibles. Now with larger buffers, better support for packet units, pictures, much more. Virtually any speed ASCII, BAUDOT, CW. Text entry via built-in screen editor! Adjustable split screen display. Instand mode/speed change. Hardcopy, diskcopy, break-in buffer, select calling, text file transfer, customizable full screen logging, 24 programmable 1000 character messages. Ideal for MARS and traffic handling. Requires 256k PC or AT compatible, serial port, RS-232C TU. $65. Send call letters (including MARS) with order. David A. Rice, KC2HO, 144 N Putt Corners Rd, New Paltz, NY 12561.

LET THE GOVERNMENT FINANCE your small business. Grants/loans to $500,000. Free recorded message: 707-448-0270. (KH5).

WANTED: Old, odd and unique bugs (speed-keys). Please help. Smiley, WB4EDB, PO Box 5150, Fredericksburg, VA 22403. (703) 373-0996.

BEAT THE COLD! Melbourne, FL QTH: 4:2:2 (3100 sq/ft) house, 1.6+ acres, 220V wired shack with coax races built in, RHON/25 pad, workshop, pool, sprinkler system, in the country no restrictions, many tall pines and oaks, easy commute to Cape Canaveral, close to fishing/beaches/shopping, SAE for details and photos, $209,000. N0DH/4.

AVANTEK ATF10135, $12.00, MMIC's, P.C. board, SASE: WA3IAC, 7148 Montague St, Philadelphia, PA 19135.

"HAMLOG" COMPUTER PROGRAM. Full features, 18 modules. Autologs, 7-band WAS/DXCC. Apple, IBM, CP/M, KAYPRO, Tandy, C128 $24.95. HR-KA1AWH, POB 2015, Peabody, MA 01960.

HF PACKET IMPROVEMENT—Digicom 64 revision "A" circuit now available. Board plugs directly into cassette port or remote mount via cable, both connectors included. Power derived from computer. Uses 7910 chip—no alignment required. Switch allows HF or VHF operation. Order Kit #154 for $49.95 or Assembly #k154 for $79.95, both include FREE DISK. Add $3.50 s/h, CA add 6.25% tax. A & A Engineering, 2521 W. LaPalma #K, Anaheim, CA 92801. (714) 952-2114. MC or VISA accepted.

IMRA International Mission Radio Association helps missionaries. Equipment loaned. Weekday net, 14.280 MHz, 1-3 PM Eastern. Nine hundred Amateurs in 40 countries. Rev. Thomas Sable, S.J., University of Scranton, Scranton, PA 18510.

HOW TO DESIGN AND BUILD YOUR OWN COMPUTERIZED PRODUCT! New manual makes it easy! Simple step-by-step details from idea to finished product. Includes introduction to design, choosing the proper CPU, comparison between many CPUs & microcontrollers, circuit layout, sample schematics, programming, debugging, hints, tips, and more! $8.00 complete. Satisfaction guaranteed. Details free. TRAVCO, 107 E. Vallette, Suite 1301G, Elmhurst, Illinois 60126.

HAM QTH. Hilltop woodland. Approx. 1500 ft. above sea level. 17.8 acres. Located in NH near Mass. border; perc-tested for single residence, solar exposure, privacy, access road is town-owned, owner-maintained. Phone and power lines within approx. 2500 ft. $79,000. Call or write for details. (603) 547-2053. C.W. Farr, Broker, Rt 1, Box 223, Greenfield, NH 03047.

INTERESTED IN PUBLIC SERVICE? Join your Local Radio Emergency Associated Communications Team. In Pennsylvania call (717) 938-6943 or write REACT, 1160 Old Trail Rd, Etters, PA 17319.

RUBBER STAMPS: 3 lines $5.00 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

ELECTRON TUBES: Receiving, transmitting, microwave... all types available. Large stock. Next day delivery, most cases. DAILY ELECTRONICS, PO Box 5029, Compton, CA 90224. (213) 774-1255.

RECONDITIONED TEST EQUIPMENT $1.25 for catalog. Walter, 2697 Nickel, San Pablo, CA 94806.

COMING EVENTS

Activities — "Places to go . . ."

SPECIAL REQUEST TO ALL AMATEUR RADIO PUBLICITY COORDINATORS: PLEASE INDICATE IN YOUR ANNOUNCEMENTS WHETHER OR NOT YOUR HAMFEST LOCATION, CLASSES, EXAMS, MEETINGS, FLEA MARKETS, ETC. ARE WHEELCHAIR ACCESSIBLE. THIS INFORMATION WOULD BE GREATLY APPRECIATED BY OUR BROTHER/SISTER HAMS WITH LIMITED PHYSICAL ABILITY.

MARCH 31-APRIL 1: MARYLAND: The Baltimore ARC's 1990 Greater Baltimore Hamboree and Computerfest, Maryland State Fairgrounds Exhibition Complex, Timonium. Gates open 8 AM. Admission $5 for both days. Children under 12 free. For information and space reservations contact GBH&C, PO Box 95, Timonium, MD 21093-0095 or 1-800-HAM-FEST 24 hours a day. In MD dial 301-HAM-FEST on TT phone.

APRIL 7: MINNESOTA: The Lake Region ARC's 3rd annual Hamfest, Otter Tail County Fairgrounds Hockey Arena, Hwy 59 So., Fergus Falls. 8 AM to 3 PM. Registration $3/advance; $4/door. 6' tables $4. For information call (218) 826-6274 or write Keith McKay, N0FKF, Rt 1, Box 46, Battle Lake, MN 56515.

APRIL 7: MICHIGAN: The Blossomland ARA's Hamfest, Fair Plain Jr. High School, Benton Harbor. 8-12. Admission $3. For information Paul Reissmann, WD8MWT, B.A.R.A., PO Box 175, St. Joseph, MI 49085 or call (616) 429-6230.

APRIL 7: WASHINGTON: The 12th annual Inland Empire and Eastern Washington Section Hamfest, Spokane Youth Sports Bingo Hall, E. 2230 Sprague Ave, Spokane.

APRIL 7: PENNSYLVANIA: AARG Hamfest and Computer Show, Lebanon Fairgrounds, Lebanon. Admission $4. Tailgating $3/space. Handi accessible. For information contact Homer, WA3YMU, (717) 345-3780, Ron, WB3HNX (717) 345-8667, Paul, WB3HEC, (717) 566-2606.

APRIL 8: NORTH CAROLINA: The Raleigh ARS is sponsoring its 18th annual Hamfest and NCS ARRL Convention, Jim Graham Building, NC State Fairgrounds, Hillsboro St, Raleigh. Admission $4 to April 2. For information contact Rollin Ransom, NF4P (919) 269-4406.

APRIL 14-15: TEXAS: The Key City ARC' Swapfest and ARRL Convention, Abilene Civic Center, Pine St.8 AM to 5 PM and 9 AM to 4 PM Sunday. Admission $5 advance; $6/door. Tables $2. Send pre-registration to KCARC, PO Box 2722, Abilene, TX 79604 or contact Bill Jones, N5DOX (915) 698-4606.

APRIL 15: MASSACHUSETTS: Tailgate electronics, computer and Amateur Radio flea market, sponsored by the MIT Radio Society and the MIT Electronics Research Society, Albany and Main St, Cambridge. Admission $1.50. Free off street parking for 1000 buyers. For space reservations or info call (617) 253-3776. Mail advance reservations before April 5 to Richard Brezina, 3 Ames St, Cambridge, MA 02139.

APRIL 21: OKLAHOMA: The Lawton-Fort Sill ARC's 43rd annual Hamfest, County Fairgrounds, Lawton. 8 AM to 5 PM. For information contact Claude R. Matchette, 3411 NW Atlanta Ave, Lawton, OK 73505. (405) 357-5870.

APRIL 21: ARKANSAS: ARRL All Arkansas Ham Convention, North Little Rock Community Center. Sponsored by the C.A.R.E.N. Radio Club. Admission $3. For information contact Bob Hancock, KB5IDB (501) 771-2617 or Leon Schlosser, N5GYW (501) 835-4004 evenings.

APRIL 22: MASSACHUSETTS: Wellesley ARS Hamfest, Wellesley Senior High School parking lot, 50 State Street, Wellesley. 9 AM to 2 PM. Admission $2. Contact Gerry Driscoll, NV1T (617) 444-2686.

APRIL 22: CONNECTICUT: The 7th annual Southington ARA's Flea Market, Southington HS Cafeteria, 720 Pleasant Street, Southington. 9 AM to 1 PM. Admission $3. For information Chet, KA1ILH, 628-9346, 5 to 9 PM.

APRIL 22: ILLINOIS: The Moultrie ARK's 27th annual Hamfest, Cadwell Road, Sullivan. Tickets $4. For information, tables, space call or write Ralph Zancha, WC9V, 502 E. State St. Lovington, IL 61937. (217) 873-5287 evenings.

APRIL 27: DAYTON, OHIO: The Dayton-Cincinnati Chapter of the Quarter Century Wireless Association announces its annual QCWA Banquet, Neil's Heritage House. COD bar 6:30. Banquet 7:30. Tickets $14, reservations required. For tickets or information contact Bob Dingle, KA4LAU, 657 Dell Ridge Drive, Dayton, OH 45429. (513) 299-7114.

APRIL 27: DAYTON, OHIO: The Southwest Ohio DX Association will again host the DX dinner at the Stouffer Center Plaza Hotel. Advance tickets only $23.00 per person. SASE to with check payable in US funds to SWODXA to Scott Lehman, N9AG, PO Box 803, Greenville, OH 45331.

APRIL 27: DAYTON, OHIO: The 21st annual B*A*S*H*, Conference Center (Madison Room) of the HARA Arena. Starts 7 PM. No admission charge. Hot food available. Free entertainment. Also a hot-air balloon launch Saturday, April 28 at 6:45 AM (weather permitting) conducting aeronautical QSO's. Frequencies announced at B*A*S*H*. Special event cards will be issued. Sponsored by the Miami Valley FM Association, PO Box 263, Dayton, OH 45401.

APRIL 29: NEW HAMPSHIRE: "Dayton East", Contoocook (near Concord). Starts 7 AM. Buyers $1. Sellers $5. For information K1OPQ (603) 746-5090 or PKT BBS ⁹ WA1WOK-2.

MAY 4-6: ARIZONA: The Cochise ARA's annual Hamfest, club training facility, Moson Road, Sierra Vista. Free tailgating. Handi facilities. For information contact N7INK (602) 378-3155 after 6 PM or write CARA, PO Box 1855, Sierra Vista, AZ 85636.

MAY 5: WISCONSIN: The Ozaukee Radio Club's 12th annual Cedarburg Swapfest, Circle B Recreation Center. 8AM to 1 PM. Admission $2/advance; $3/door. For tickets, table reservations, maps, information send business SASE to ORC Swap, N5415 Crystal Springs Ct, Fredonia, WI 53021. (414) 692-2329.

MAY 20: ILLINOIS: The annual Hamfest sponsored by the Kankakee Area Radio Society, Will County Fairgrounds, Peotone. 8 AM to 2 PM. Indoor flea market and exhibitor tables (limited). Large outdoor flea market. ARRL booth, free parking. Food and drink available. Admission $2.50 advance, $3.00 at the door. Setup kMay 20, 6-8AM. Talkin on 146.34/94. More information from KARS, c/o Frank DalCanton, KA9PWW, RR1, Box 361, Chebanse, IL 60922. Tel (815) 932-6703 after 4 PM CST or (815) 937-2452 before 4 PM CST.

OPERATING EVENTS

"Things to do . . ."

The Tipton ARS will operate special event station KC7YL from Fort Pillow State Park in Lauderdale County, Tennessee. 1600Z April 7 to 1900Z april 8. Voice: 28.450, 21.350, 14.325. CW: 28.200 and 21.150. For an attractive certificate send QSL and large SASE to WB4DPL, PO Box 402, Covington, TN 38019.

The Cornish Radio Amateur Club of England is sponsoring International Marconi Day to celebrate Marconi's birthday. 0000Z-2400Z April 21 in the General bands. Work 10 stations and send log to CRAC, PO Box 100, Truro, TX1 1RX, Cornwall, UK.

The Warminster ARC's 2nd annual DXpedition to Delaware, April 8, operating WA3DFU/3. Freq: 7.275, 14.275, 28.375. CW on request. QSL with SASE to Warminster ARC, Box 113, Warminster, PA 18974.

The Great River ARC, Dubuque, IA, will operate special event station WK90 (N9GBY) April 28 at the 36th annual U.S. Boy Scout Pilgrimage, 10-4 local time, lower General band and Novice 10m band, SSB. No SAE, QSL required for certificate.

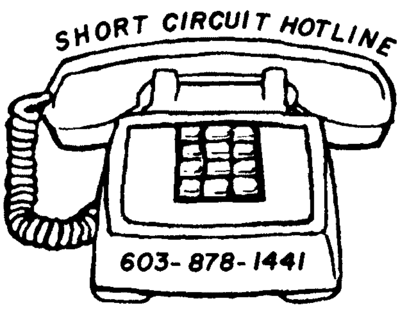
Clairemont Repeater Association will operate "Sam's Day" in honor of the 199th birthday of Samuel F.B. Morse. Control op W6FZZ, Samuel Morse 3d. 1800 UTC to 2400 UTC Saturday April 28. 28.350 and 21.300 phone and 14.050 CW. QSL to CLARA, PO Box 7675, Huntington Beach, CA 92615.

THROUGHOUT 1990 the Major Armstrong Memorial Amateur Radio Club (MAMARC) will sponsor events commemorating Major Edwin Howard Armstrong's achievements in the field of radio broadcasting. The club is seeking other Amateur operators around the world who are willing to research Major Armstrong's accomplishments and become official MAMARC special events stations. Major Armstrong was a pioneer responsible for the creation of Wideband FM and the inventor of the superhetrody

Monthly Ham Exams. The MIT UHF Repeater Association and the MIT Radio Society offer monthly ham exams, all classes Novice to extra: next-to-last Wednesday of each month, (April 18) 7:30 pm, MIT Room 1-150, 77 Mass Avenue, Cambridge, MA. reservation requested a couple of days in advance, walk-ins welcome; call the shack (617) 253-3776, or Nich Altenbernd (617) 437-0320. Exam fee $4.95. Bring copies of your current license (if any) and Certificates of Completion (if any), two forms of picture ID, and a completed form 810, available from the FCC: (617) 770-4023.

LAUREL ARC monthly (except December) Amateur exam sessions for all license classes. No fee is charged. Pre-registration is required. Call (301) 725-1212, Maryland Radio Center, 8576 Laureldale Drive, Laurel, MD 20707.

AMATEUR RADIO CLASSES: For those people interested in obtaining a Novice (basic level) Ham license or upgrading to Tech/General, the Chelsea Civil Defense, in cooperation with QRA Radio Club, will sponsor Amateur Radio Communications classes evenings at Chelsea High School starting MARCH 7, 1989. For more information write Frank Masucci, K1BPN, 136 Grove Street, Chelsea, MA 02150. Please enclose your telephone number.



MOVING?

Let circulation know 60 days in advance.

Ham Radio
Greenville, NH 03048-0498
603-878-1441

By Tom McMullen, W1SL

ELEMENTARY ELECTRONICS: INDUCTORS AND AC

In past issues I've looked at resistors, capacitors, and magnetic fields around wires and simple coils. I explored them using direct current (DC) as the energy source. Life in the DC world is somewhat simple. Current either flows or it doesn't, and when it flows, it goes from one power source terminal to the other in a straightforward manner Along the way, it can be restricted by resistors, smoothed by capacitors, divided, combined, and so on. It's all just "plain vanilla" electronic theory. When you enter the world of alternating current (AC), things get a bit more complex. Most of the simple rules apply, but you have to keep one basic rule in mind — alternating current flows from one power source terminal to the other for a short time, *and then turns around and goes back!* Secondary to this is the fact that the voltage isn't constant, but varies from zero volts to some nonzero value (either positive or negative) and then returns to zero twice during each cycle.

Waveforms

The figures you see on an oscilloscope are generally voltage waveforms, so it's a bit strange to call them AC waveforms. However, if you remember that the voltage you're looking at was probably derived from alternating current flowing through a resistance, it makes sense that the oscilloscope shows a representation of an AC waveform.

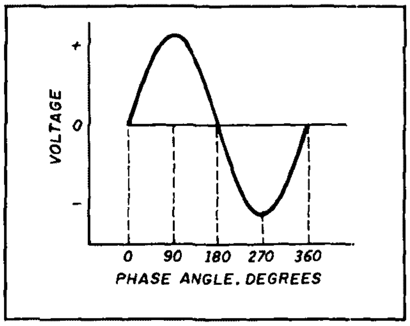
A basic AC waveform, like the one provided by most household electric service, is shown in Figure 1. Household AC is generated by mechanical means — a generator that forces a massive winding of wire through a magnetic field. Waveforms generated by electronic circuits (oscillators) can be made with the same shape and cycle period, or with an almost unlimited variety of shapes and cycles.

Current and voltage relationships

It doesn't matter if the waveform you're examining is made by a mechanical generator or is created by electronic oscillators; the relationship between the voltage and current is what's important.

At the start of each cycle, there's no voltage and no current flow; this is the zero (0) point. As the voltage builds up, current increases. In a circuit with only pure resistance, the current and voltage stay in unison throughout the cycle; when the voltage reaches its peak, so does the current. The complete cycle is said to take place in 360 degrees, and is related directly to the mechanical generator concept where a coil (armature) rotates about its axis while in a magnetic field. Portions of a cycle can be described conveniently by referring to the "degrees" (of rotation) along the waveform. This degree concept also comes in handy when determining how the current is doing compared with the voltage (more on this later).

FIGURE 1



The waveform of common household AC power is a sine wave with equal positive and negative peaks during a cycle. The degree references are explained in the text.

The waveform demonstrated in **Figure 1** is called a sine wave. You can generate one on graph paper by plotting the sine of the angle, as shown in **Figure 2**. Not all AC waveforms are as neat as this, however. They can be distorted by amplifiers, capacitors, and inductors, or they may not have been sine waves to start with. Square, triangle, and sawtooth waves, in addition to random audio (speech) waveforms, are also types of AC. But the easiest way to understand how AC behaves in a circuit is to stick with sine waves.

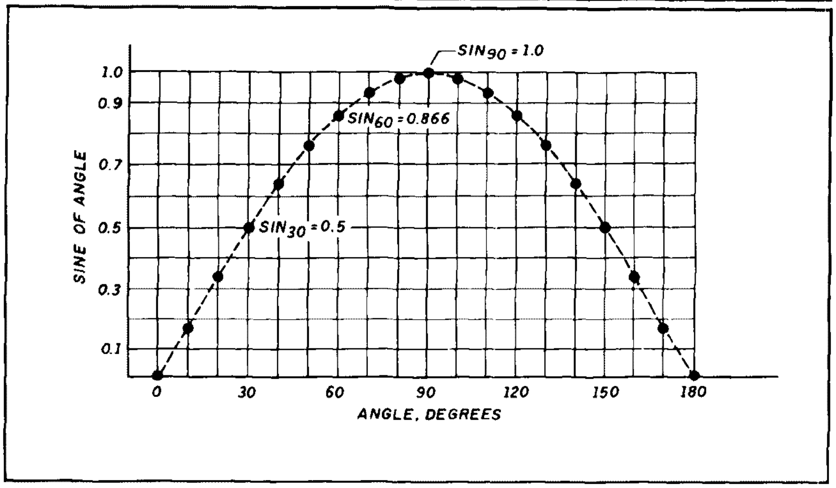
Phase shift

In simple resistive circuits, the current flow stays exactly in time with the voltage buildup or decay. When the voltage is at maximum, so is the current flow. As voltage changes, so does the current — in exact proportion. Under these conditions, the voltage and current are said to be "in phase" with each other. However, AC really spends most of its life in a world that's full of reactance (I'll talk about reactance shortly). Both inductors (coils) and capacitors have reactance, but I'll just look at inductors this time and save capacitors for a future issue.

In a past discussion of DC flowing in an inductor, I noted that the current flow created a magnetic field which not only used and stored energy, but created a counter force (called counter electromagnetic force or counter emf) that opposed the current flow during the buildup. When the current flow stopped, the magnetic field collapsed and the energy was released — returned to the circuit in the form of electron flow in the same direction as the original current flow!

The same thing happens when AC flows through a coil, but because of the changing nature of AC, the effects are more noticeable. As the voltage builds up through the 0 to 90-degree portion of the waveform, the current gets behind. Remember, it's current (electron movement) that does the work in a circuit. So while the voltage gets from one end of a coil to the other quickly, the current is busy building up a magnetic field which further slows the current buildup because it creates a

FIGURE 2



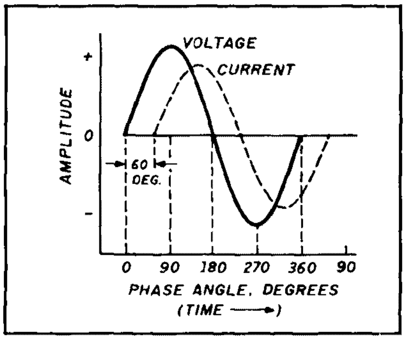
A sine wave can be drawn on quadrille paper by plotting the sine ot the angle.

counter emf that's pushing back.* As a result, current lags behind voltage, as shown in **Figure 3**. The voltage and current are now "out of phase" with each other. Remember when looking at **Figure 3** that time progresses to the right. Although the current is delayed in time with reference to the voltage, it's far easier to express the delay in degrees. This is because the time for one cycle varies, depending on the frequency. For household power, a cycle occurs 60 times per second, or one cycle equals 0.016666 second. At a frequency in the Amateur 80-meter band, a cycle might occur 3.6 million times per second, or approximately 0.0000002 second for one cycle. If you use degrees, you don't have to worry about how small the fraction of a second is.

The number of degrees the current lags the voltage depends on the inductance value and the frequency of the AC. If the inductance is very small, say ten or 12 turns of wire on a plastic form, and the AC frequency is very low, like 60 cycles per second (60 Hz), the current lag is undetectable. However, with those same ten turns of wire and a frequency of 500 million cycles per second (500 MHz), the inductance has a very great effect on the current — so great that no alternating current can flow through the coil. (This is the principle behind an RF choke.)

Aha, you caught me trying to sneak a term past you! Inductance. Let's

FIGURE 3



An inductor (coil) causes the current flow to lag behind the voltage, 60 degrees in this example. The voltage and current amplitudes are not drawn to scale.

define it before continuing. Inductance is the tendency of a wire or coil to oppose any change of current flow through it. It does this by using the magnetic field to first oppose a current increase, and then release the energy stored to oppose the current decrease. The unit of inductance is the henry, abbreviated H; smaller units are the millihenry (mH) and microhenry (μH). The units do not take on the "ies" suffix for plural; it's millihenrys, not millihenries.

Inductance is inductance, and it doesn't matter whether you're trying to make AC or DC flow through it — it opposes both. So you need a way to relate its opposition to the frequency involved, because you've seen that the inductance of a ten-turn coil doesn't bother 60 Hz, but completely shuts out 500 MHz. The property that relates inductance and frequency is called inductive reactance, X_L. Because inductive reactance tends to limit current flow in a circuit, it is expressed in ohms.

X_L can be calculated at the frequency of interest. For example, a coil with an inductance of 1 mH (0.001 H) at a frequency of 4 MHz (4,000,000 Hz) will have an inductive reactance (X_L) of:

$$X_L = 2\pi fL$$

where f = frequency in Hz
L = inductance in henrys (H)

thus:

$$X_L = 6.28 \times 4{,}000{,}000 \times 0.001$$
$$= 25{,}120 \text{ ohms}$$

Many circuits have resistance as well as inductive reactance, so it's necessary to find some term that combines the two for ease of calculation. This term is impedance. The symbol for impedance is Z.

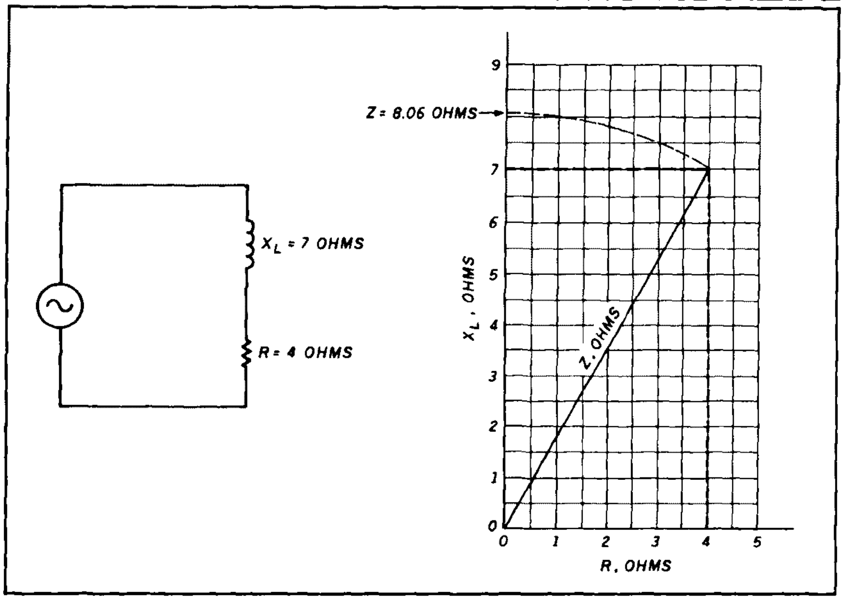
Here's where some of you will get cold chills, blurred vision, and sweaty palms — but hang in there, what follows isn't life threatening (unless you're in the middle of final exams). Impedance is the vector sum of the inductive reactance and the resistance in a circuit.

Oh, no, not vectors!

I know, I hated vectors too. But like much of life, once you understand them, they're not half bad. To simplify the vector concept a bit, look at a vector as the result of two forces working on the same object. Imagine that you're trying to move a ping-pong ball from your end of the table to the other by blowing on it. At the same time, another person is trying to move the ball from one side to the other in the same way. As a result, the ball takes a path at an angle to both the end and the side of the table. The exact path depends on how hard each of you blows. The amazing thing about vectors is that they add the forces involved. If the other person is at the opposite end of the table instead of the side, the ball might follow a straight path but wouldn't get as far. The opposing negative force counteracts your positive force, and the result is the sum of the two. Thus, if two people were at the other end of the table, the vector could be even more negative, and the ball would go off the table at your end. This

* Don't misunderstand what's being delayed here. The counter emf opposes the number of electrons flowing, but doesn't change the speed of those electrons.

FIGURE 4



A vector diagram can illustrate the impedance of a circuit with both reactance and resistance.

ability to sum positive and negative forces is important; you'll see more of it in a future column.

Back to the electronic world, before I run out of space. **Figure 4** shows an example of a vector used to determine the impedance (Z) of a circuit that has both inductive reactance and resistance. The inductive reactance (X_L) is plotted vertically, and resistance (R) is plotted horizontally. You can do this on ordinary quadrille paper. The hypotenuse represents the impedance, and for this example is approximately 8 ohms.

Impedance can also be calculated by rearranging the formula for finding the hypotenuse in the Pythagorean theorem: "The square of the hypotenuse value of a right triangle is equal to the sum of the squares of the values of the other two sides." Or,

$$Z^2 = R^2 + X_L^2$$

Rearranged:

$$Z = \sqrt{R^2 + X_L^2}$$

Using the values from **Figure 4**:

$$Z = \sqrt{4^2 + 7^2} = \sqrt{16 + 49} = \sqrt{65} = 8.06 \text{ ohms}$$

The vectors can also be used to determine how much the current lags the voltage in a circuit. This is only one of the intriguing things yet to be explored. Stay tuned!

Conclusion

You've seen how inductance affects alternating current, causing current flow to lag behind the voltage. This is *important* to engineers when designing power, audio, and RF circuitry. It's also important to you for understanding how a properly designed circuit can provide maximum power transfer between stages of a transmitter, or between your transmitter and an antenna. You'll need another term to help do this — capacitance — which we'll look at in the future.

DX Forecaster

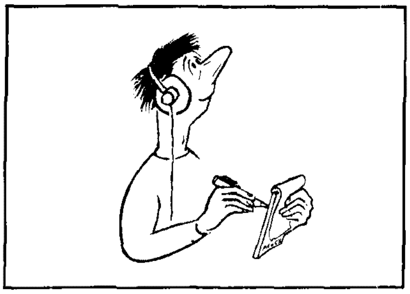
By Garth Stonehocker, KØRYW

DX PROPAGATION SOFTWARE

More and more ham shacks sport computers these days. The popular PCs with up to 640K of on-board computing storage give power that a few years ago was available only in institutions, businesses, and government laboratories. Now Amateurs have new help in logging, operating, and performing computer-aided design of antennas and equipment. Much of the software to accomplish these tasks has been published or advertised for sale in Amateur Radio magazines over the last few years. Today, with the different software packages available, PCs can even be used as aids in propagation prediction.

The early use of lowest and highest usable frequencies (LUF/MUF) predictions was once limited to those who received Amateur Radio magazines or worked where these propagation programs were part of their business activities. (The Institute for Telecommunciations Science, with ITS-78 and IONCAP, is one such place.) In December 1982 Navy researchers developed a propagation program called MINIMUF-3. Written in BASIC, this program was made available to Radio Amateurs in the December 1982 issue of *QST*.[1] It calculated the MUF for each hour of the day for any month and the sunspot number for station-to-station paths, by latitude and longitude. The appearance of MINIMUF-3 led to a proliferation of modifications of the MINIMUF program. In most cases the output was changed to do a specific task or produce a more interesting display. *Ham Radio* published such a display modification in February 1987.[2] It showed the MUF as pseudo-concentric contours around a transmitter.

Advertised propagation prediction programs usually cost between $25 and $50 and support Commodore, Macintosh, and IBM compatibles. Several of these commercial offerings display the MUF/LUFs on a map of the world. Most have grayline propagation, sunrise/sunset times, and antenna bearings. Typical of these are The DX



Helper by W7HR, The DX Edge,* and BANDAID/MUFMAP by Heath. Some programs, like MINIPROP by W6EL, include signal level information. The ultimate software for signal and noise levels is W1FN's IONOSOUND. The program determines the signal to noise and reliability of each propagating mode (E, F, or mixed) over the path, at frequencies you select as being most useful for DX. The display simulates an oblique chirp sounder showing the modes. Combining a MUF program with IONOSOUND gives you capabilities similar to those of the ITS-78 and IONCAP programs — with the exception of world noise and antenna patterns. So you can bring the power of the mainframes of a few years ago to your ham shack PC and use it to help find DX openings.

How do you know how "good" these programs are? Play the beacon game. Pick a beacon path close to your DXing path of interest, if possible. If just one beacon frequency is transmitted, the whole diurnal MUF curve for that path will be hard to define. In any case, monitor the frequency until the signal disappears in the afternoon, defining a time point on the curve. If the signal disappears at a later hour the next day, the MUF was probably higher for that time period. Monitor the signal for several days to build up the statistical average. Use the same technique in the morning as the signal arrives. The morning points shouldn't show as much scatter in time as the afternoon's MUF downward curve.

Last minute forecast

The second and third weeks are the days of high MUFs (high solar flux). They may decrease signal strengths but lengthen the nearly daily openings on the higher HF bands, 10 to 20 meters. This month's evening one long hop transequatorial openings to the south could come on the 11th and 19th. Other ionospheric disturbed conditions may exist near the 1st and 28th. The low band east-west paths may have weak signals and QSB caused by the two latter disturbances. The low bands will be best at times of low disturbance.

The perigee of the moon's orbit (for moonbounce DX) is on the 26th, with the moon at full phase on the 10th. There will be a short meteor shower, the Lyrid, on April 20th to 22nd with a rate of five per hour — hardly much help for meteor-scatter DX. But a bigger shower, the Aquarid, starts before the end of April, peaks on the 5th of May, and ends in mid-May. Its rate is 10 to 30 per hour.

Band-by-band summary

Ten, 12, 15, 17, and 20 meters will be open from morning until late evening most days to most areas of the world. The higher bands open closer to local noon and mainly to southern countries. One long hop openings will probably occur in late evening during high solar flux periods with a mildly disturbed geomagnetic field. The lower of these bands work mainly to the north and south in the evenings.

Thirty, 40, 80, and 160 meters will be open mainly at nighttime. The highest band, 30, is open near dawn and dusk on northern paths. The other band's signals are so absorbed during the day at this high sunspot number (SSN) that DXing is relatively rare. Good nighttime signals are present this time of year with low thunderstorm activity. So, except for an occasional weather frontal passage by your QTH, these conditions should make DXing a joy. Disturbed conditions from the high SSN solar activity may bring weak and fluttery signals from interesting places at night on these bands.

* Available in C-64 and MS-DOS versions from the HAM RADIO Bookstore for $34.95 plus $3.75 shipping and handling.

REFERENCE

1. Robert B. Rose, K6GKU, "MINIMUF: A Simplified MUF-Prediction Program for Microcomputers," *QST*, December 1982, page 36.
2. Henry G. Elwell, Jr., N4UH, "360-Degree MINIMUF Propagation Prediction," *Ham Radio*, February 1987, page 25.

APRIL

WESTERN USA

GMT	PST	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	4:00	12	20	12	10	10	10	10*	10
0100	5:00	12	20	15	10	10	10	10*	10
0200	6:00	12	30	15	10	10	10*	10*	10
0300	7:00	12	30	15	10	12	10*	10*	12
0400	8:00	12	17	17	12*	12	10	10	12
0500	9:00	15	17	15	12	17	10	10	12
0600	10:00	20	20	12	15	17	10	10	15
0700	11:00	20	20	12	15	20	10	10	17
0800	12:00	20	20	15	15	20	10	10	17
0900	1:00	20	20	15	15	20	12	12*	17
1000	2:00	20	30*	15	17	20	12	12	20
1100	3:00	30	17	15	15	20	15	15	20
1200	4:00	17	15	10	15	20	15	15	20
1300	5:00	15	12	10	12	20	15	15	20
1400	6:00	15	12	10	10	20	17	17	20
1500	7:00	20	10	10	10	20	17*	15	20
1600	8:00	20	10	10	10	15	17	15	20
1700	9:00	20	12	10*	10	12	15	17	20
1800	10:00	20	12	10	10	10	12	17	17
1900	11:00	20	12	10	10	10	10	12	15
2000	12:00	30	12	10	10	10	10	10	12
2100	1:00	17	15	10	10*	10	10	10	12
2200	2:00	15	15	10	10*	10	10	10	12*
2300	3:00	12	17	10	10	10	10	10	10

MID USA

GMT	MST	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN	CST
0000	5:00	12	20	12	10	10	10	10	10	6:00
0100	6:00	12	20	15	10	10	10	10*	10	7:00
0200	7:00	15	30	15	10	12	10*	10*	12	8:00
0300	8:00	15	30	15	10	12	10*	10	12	9:00
0400	8:00	20	17	17	12	15	10	10	12	10:00
0500	10:00	20	15	15	12	17	10	10	15	11:00
0600	11:00	20	20	12	15	17	10	10	17	12:00
0700	12:00	20	20	15	15	20	10	10	20	1:00
0800	1:00	20	30*	15	15	20	12	12	20	2:00
0900	2:00	30	30	15	17	20	12	12	20	3:00
1000	3:00	30	17	15	15	20	15	12	20	4:00
1100	4:00	17	15	12	15	20	15	15	30	5:00
1200	5:00	15	12	10	12	20	15	17	17	6:00
1300	6:00	15	12	10	12	20	17	15	17	7:00
1400	7:00	20	12	10	10	20	15	15	20	8:00
1500	8:00	20	10	10	10	20	15	15	20	9:00
1600	9:00	20	10	10	10	15	15	15	20	10:00
1700	10:00	20	12	10*	10	12	17	17	20	11:00
1800	11:00	20	12	10	10	10	12	17	17	12:00
1900	12:00	20	12	10	10	10	10	12	15	1:00
2000	1:00	20	12	10	10	10	10	10	12	2:00
2100	2:00	30	15	10	10*	10	10	10	12	3:00
2200	3:00	20	15	10	10	10	10	10	12	4:00
2300	4:00	15	15	10	10	10	10	10	12	5:00

EASTERN USA

GMT	EST	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE CARIBBEAN S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	7:00	15	20	12	10	10	10	10*	12
0100	8:00	15	20	15	10	10	10*	10	12
0200	9:00	20	20	15	10	12	10	10	15
0300	10:00	20	30	17	10	15	10	10	17
0400	11:00	20	30	17	12	17	10	10	20
0500	12:00	20	20	15	12	20	10	10	20
0600	1:00	30	30*	15	15	20	12	12	20
0700	2:00	30	30	15	15	20	12	12	20
0800	3:00	17	17	17	15	20	12	12	20
0900	4:00	15	15	12	17	20	15	15	30
1000	5:00	12	15	10	15	20	15	15	17*
1100	6:00	12	12	10	12	20	17	17	17
1200	7:00	12	12	10	10	20	15	15	17
1300	8:00	12	12	10	10	20	15	15	20
1400	9:00	12	10	10	10	20	15	15	20
1500	10:00	15	10	10	10	15	17	15	20
1600	11:00	17	10	10*	10	12	17	17	20
1700	12:00	20	12	10*	10	12	12	17	20
1800	1:00	20	12	10	10	10	10	17	17
1900	2:00	20	12	10	10	10	10	12*	15
2000	3:00	20	12	10	10	10	10	10	12
2100	4:00	30	15	10	10	10	10	10	12
2200	5:00	20	17	10	10	10	10	10	12
2300	6:00	15	20*	12	10	10	10	10	12

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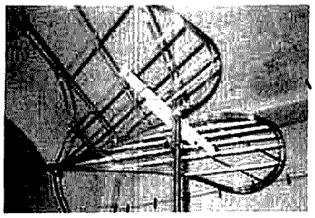
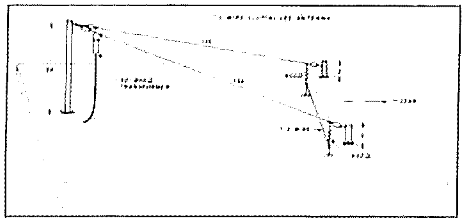


K3OQF, page 11



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FEATURES

DEPARTMENTS

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Backscatter

FOR SALE: THE 20-METER AMATEUR BAND. BIDS START AT $2,000,000,000. GET YOUR AUCTION NUMBER NOW!

How's that for a headline! Think I'm kidding? Read on and learn the exciting details of this fascinating opportunity.

The National Telecommunications and Information Administration (NTIA) and the FCC are responsible for the management and allocation of the radio spectrum for the federal government. On December 8, 1989, the NTIA filed a long and involved Notice of Inquiry (NOI) which was published in the Federal Register. It dealt with a number of important subjects.

This NOI covered the following areas: economic impact of spectrum management, accommodation of new technologies, block allocations, separate processes for government versus private sector allocations, sharing between government and private services, private use of government frequencies, international conference preparation, unmet spectrum demands, definitions and standards, usage measurement of allocated bands, and unlicensed device interference. While this is a very broad list, it covers a number of very important areas of concern to Radio Amateurs.

Of special interest are the NTIA's thoughts on auctioning off radio spectrum to generate *revenue for the federal government*. Huh???? Who's kidding who! That's one of the best jokes I've heard in years. Ostensibly, revenues raised by auction would go to reduce the federal deficit. However, we all know that the reduction would be a drop in the bucket compared with the real value of the spectrum that they're proposing to sell. And what about future value? Would this be a one-time auction, or would we be bidding for a yearly lease on the frequencies?

On February 23rd the ARRL filed a rather lengthy reply to the NTIA's NOI petition. They did a very complete job of addressing the concerns of Amateur Radio operators. Stating that the League has been involved in spectrum management for 76 years, they commented: "Just as public policy requires that there be public parks and other open space for the nourishment of the human spirit, so must there be 'open spaces' in the radio spectrum where the public can pursue personal, noncommercial ends. This is the essence of the Amateur Radio Service."[1]

There are certain parts of our American culture that the government holds in trust for its citizens. National parks, monuments, and the treasures of the Smithsonian Institution are all kept and preserved for us by the government. Imagine the outcry if the Smithsonian were to auction off the Hope Diamond. It would generate millions — maybe even billions. But at what cost? A number of groups are now fighting the Parks and Forest Services' attempts to auction off mineral and timber rights in our national forests and parks. Sure, this will raise huge sums of money — but at the irrevocable loss of a national treasure.

A similar parallel can be made to selling frequencies — once they're gone, well, it's all over, they're gone!

Over the years, in times of disaster, the Amateur Radio community has provided continuity of communications for government and non-government relief agents and the public, when normal communications links have been disrupted or completely severed. To quote the ARRL once again, "The Amateur Radio Service, through participation by thousands of dedicated volunteers, provides a resource that could never be matched by commercial services...A pure marketplace approach to spectrum management would work to the serious detriment of Radio Amateurs, and thus to the public interest."

What the League doesn't say in its statement, because it's "playing" politics and can't, is that this is a simply ridiculous proposal. Sooner or later our government has to come to grips with the deficit and implement a fair and equitable plan to bring it under control. Auctioning off spectrum *is not* the way to do it. At best it is a stop-gap measure that creates as many problems as it resolves. In the end, we all end up as losers!

Even if there were to be an exclusion for Amateur Radio which said we wouldn't have to bid competitively for "our bands," this would still be an ill-conceived and poorly thought out proposal. However, if we don't assert ourselves and let the NTIA, FCC, and our elected representatives know that we feel this is a stupid proposal, we may have to live with the consequences of our inaction.

We urge you to stay current with this issue. It could result in major changes in Amateur Radio as we know it.

de NX1G

1. "ARRL comments in NTIA Spectrum Management Inquiry," *The ARRL Letter*, Volume 9, No. 5, March 16, 1990, page 1.

Editor's Notes

Well, the April Fool's joke was on me! As many of you noticed, the article "Trap Dipole for 12 and 17 Meters" listed on the cover of our April 1990 issue was nowhere to be found inside the magazine. I'd like to be able to say I was just checking to see if anyone reads the cover, but I must confess that I simply forgot to include it. For those who have been eagerly awaiting its arrival, the piece (by Gary Nichols, KD9SV) appears on page 22 of this issue. My apologies to Gary and to all who've had to wait an extra month to read about this interesting antenna.

Terry Northup, KA1STC

Comments

Repeater etiquette

Dear HR

In the January 1990 issue, K6BAZ laments the fact that only 1 percent of the 440 repeaters in southern California are open to the licensed public. While I'm not sure that the other 99 percent are truly closed, I think your readers will gain some insight when they ask themselves the following questions.

1. Do you feel you have a right to use any repeater you choose, anytime you choose?
2. Do I have a right to come into your shack and operate your equipment any time I like?
3. Do you financially, or in other ways, support the repeater(s) you use?
4. Do you and your friends talk to the other users, or do you consider the repeater your own private intercom?
5. Do you complain on the air (or otherwise) when the repeater is down or sick?
6. Is tone squelch on a repeater receiver only to deny access, or could it be used for legitimate technical reasons?

I ask you to consider your answers to these questions carefully. Then, try to envision yourself as a repeater owner and how you would want others to use *your* repeater. I think you'll soon understand why some repeaters are closed, some are run with an iron fist, and why some go off the air.

Closed repeaters are a result of a lack of understanding and common sense operating. Please take the time to think about your own behavior, and perhaps your favorite repeater won't be closed.

Paul M. Alberghini, WA1KAH
Cumberland Center, Maine

Nice to hear...

Dear HR

I wanted you to know how much I enjoy *Ham Radio*. Each issue seems to get better. It doesn't surprise me that your number of subscribers is growing.

Richard Steck, W9RS,
Lake Forest, Illinois

Kudos from a pro

Dear HR

This is my first year's subscription to your fine magazine and I'm impressed not only with the content and articles, but with the quality of the printing.

I've been a pressman for some 45 years and currently run a five-unit web press. My congrats to the press crew for the excellent color and "in register" production.

Ham Radio rates a number one spot in a collage of publications for the Amateur. 73.

Dick Burnham, KA1ILR,
No. Abington, Massachusetts

No guarantees with your license

Dear HR

I wish to take exception to remarks made by N6SWA in the February 1990 "Comments" column. These comments pertained to use of the 10-meter Novice phone band by holders of General and above licensees.

I think that Mr. Gleeson is not considering the fact that those of us who hold these licenses worked to earn them, and by doing so gained the ability to operate in larger portions of the spectrum. With our incentive licensing system, it would be neither reasonable nor logical to make portions of the spectrum available exclusively to those holding the classes of license that require the least amount of study and experience. This would, in essence, penalize people for upgrading.

Mr. Gleeson should also realize that the 28.3 to 28.5 MHz band was not available to Novices until 1987. Those with General class and above licenses have always had the privilege of running higher power in that part of the band. Revoking that privilege most likely would have caused resentment among these operators, and might have been a hindrance towards having Novice enhancement accepted by the ham community.

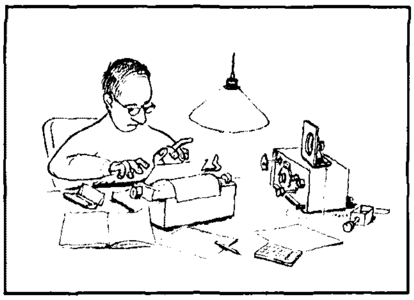
There's no guarantee of interference-free operation with an Amateur license of any class. Interference happens to all of us equally. It is absolutely ludicrous to say that it is "rude, inconsiderate, and not in the ham spirit" for Generals, Advanced, and Extras to operate legally in the Novice bands. It is, however, illegal to use more power than necessary to communicate with other stations. It is also illegal to splatter across 10 kHz of spectrum. Operators who do these illegal things, however, come in all classes of license, but are in a class of their own — that of LID.

Anyone with the Novice or Technician license who is not happy with the privileges may do one of the following three things: (1) Upgrade your license class. (2) Learn to live with it. (3) Take up a different hobby.

Stephen M. Murphy, WD8O,
East Detroit, Michigan

Correction

In a letter by Carlton D. Trotman, W3BRX, which appeared in the "Comments" section of our March issue, we inadvertently called General Curtis LeMay the *late* General LeMay. According to friends, General LeMay is living in California, where he continues to be involved in Amateur Radio. Our sincere apologies to General LeMay. Ed.



Setting The Record Straight

I would like to call your attention to an error in NU1N's article "Understanding Over-The-Horizon Radar" in the February issue of HR. The facility in Caribou, Maine which I had the privilege of commanding prior to my retirement, is not an OTH radar but a LORAN C transmitting station operated by the U.S. Coast Guard. This mistake, apparently started by Mr Helms (**Reference 9** of the article), has been perpetuated ever since. This may be due to the unique antenna in use at Caribou.

The normal LORAN C antenna is a 625-foot top-loaded monopole. This is a high Q antenna, and when used with the older vacuum tube transmitters requires frequent and tedious "pulse building" to maintain the proper pulse shape as the tubes age. Often, expensive PA tubes have to be replaced because of their inability to maintain this critical pulse shape. In the mid 1960s, the Coast Guard Electronics Engineering Center in Wildwood, New Jersey set out to design a better antenna. Results of modeling suggested that some type of conical antenna would provide the desired results. Because of lack of space at Wildwood for a full size antenna, land was purchased and an experimental station was constructed at Caribou. Development there resulted in an antenna which is an inverted wire pyramid suspended from four 700-foot towers arranged in a square 1400 feet on a side. This antenna has about double the bandwidth of the monopole and made "pulse building" a relatively infrequent and non-critical affair. When viewed from ground level, with the tower guy wires running in all directions, this antenna is hard to visualize and could be mistaken for some sort of HF directional array. Due to the expense of erecting four towers, the enormous amount of real estate involved, and the advent of new solid-state transmitters that maintain pulse shape automatically under microprocessor control, this will probably be the last of these antennas constructed for LORAN C use.

The Caribou station has since been absorbed into the LORAN network and presently transmits on rate 9960 as a secondary station in the Northeast U.S. chain and on 5930 as a master station for the Canadian East Coast chain.

There is an OTH system in Maine operated by General Electric for the Air Force. The transmitter site is at Moscow, receivers are at Colombia Falls, and the control center is at the Air National Guard facility in Bangor.

Roger D. Johnson, AD1G, CWO4, USCG (retired)

Attention antenneX Subscribers

I would like to let *Ham Radio* readers know that *antenneX*, the magazine for antenna experimenters, is no longer being published.

Unfortunately, the phone has been disconnected and the offices closed. I'm at a loss about where to direct people with questions regarding subscriptions or other items they may have ordered.

I am disappointed at this turn of events, as *antenneX* had filled a void in the market. For those of you who enjoyed the articles I wrote while I was editor of the magazine, please know that I intend to continue writing pieces on whatever comes to mind. Look for me in the pages of other Amateur Radio publications.

Richard Morrow, K5CNF, Former Editor, *antenneX*

HF Packet

HF digital communications are becoming very popular because of the availability of computers and economical all-mode modems like the PK232. The most positive thing about this is the increased use of low error rate modes like AMTOR. But there should be more concern about packet operation. As an experiment with few users, HF packet has been technically interesting and fun. As it proliferates, we need to question its fundamental design.

VHF/UHF packet operates under the assumption that almost everyone hears each other. This means that, with moderate traffic, little time is lost through retransmission to correct errors. Communication is at a high bit rate so the message bursts are short. This reduces the probability of message collision. Spectrum space at VHF/UHF is ample so there is room for several users, especially since their numbers are horizon limited.

At HF, all the above assumptions are invalid. There is no horizon limit. Ionospheric propagation favors some and ignores others. A marginal link can chew up enormous time. A trivial 3-minute QSO for RTTY or AMTOR becomes a 30-minute ordeal. The 300-baud rate is at least four times slower than VHF/UHF. Even at these rates, packet is still error prone compared with RTTY and AMTOR. This creates multiple message repeats that frustrate the link user and adjacent RTTY/AMTOR links. Yes, "gentlemen's agreement" sub-band infringement is very common on weekends.

Beyond the fun experiment, the motivation for HF packet may have been to create a nationwide emergency communications net connected to local VHF/UHF nets. There are better solutions with efficient long haul links that don't broadcast endlessly. Communications should match the medium. The rules for line-of-sight communications are not appropriate for ionospheric skip.

Hunter Harris, W1SI, *Ham Radio* Editorial Review Board

HAM-COM '90

June 8-10, 1990
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D/FW Metroplex, Texas
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Programs & Forums

Friday, June 8th:

1 PM	Gordon West's Novice/Tech License Course, WB6NOA
6 PM	Learn To Teach Code
8 PM	Slow/Fast Code Techniques

Saturday, June 9th:

9 AM	10-10 International Forum
	Contest Forum, K1AR
	Scanning & SWL, N5AUX
	Packet for Beginners
	Training to Teach Youth
10 AM	Hands On Packet
11 AM	Packet, Using DX Clusters
	VHF/UHF Technology Forum
	Traffic Handling, N5TC
	AMSAT - Microsats, N8IWJ
	VLF Radio Worldwide, N5SU
12 Noon	Packet Performance Monitor
12:30 PM	Packet, Designing Networks
1 PM	ARRL Forum
	MARS Forum
	Soc. for Preservation of AM
	Microwave UHF Forum
	YL's In Amateur Radio
3 PM	Packet Shuttle EXperiment
	Kids Only - Getting Started
	DX "Bonanza" Programs
	Wayne Green's Forum
	Amateur TV Forum, N5PIU
3:30 PM	TAPR Packet Radio,TexNet

Sunday, June 10th:

9AM	10-10 International Forum
	Contest Forum
	Gordon West's 2Mtr DX
	Protecting from Surges
	Packet for Beginners
10:00 AM	Packet, HARPS NET/ROM
11 AM	Young People in Ham Radio
	Rohn 25 Rotating Tower
	Hazardous Materials Forum
	AMSAT, Getting Started
	Satellite Uplinks/Downlinks
11:30 AM	Packet for RACES & ARES
2 PM	Skywarn School

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Ham-Com '90 is a fully recognized ARRL approved Hamfest. There will be a giant ARRL booth, an ARRL Forum and ARRL VIP's on hand.

Giant DX Lunch - Bonanza DX Program

New for Ham-Com '90 will be a giant 450 seat DX Luncheon featuring Bob Winn, W5KNE, Editor of "QRZ-DX". Tickets are $10.00 each. After lunch, we will have the DX Program with DX personalties flown in from the four corners: Dave Schmocker, KJ9I, on Mellish Reef & Willis Island, Bob Winn, W5KNE, on Sable Island, Chuck Coleman, K5LZO, on Desecheo Island, and Don Greenbaum, WB2DND, on United Arab Emirates.

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Ham-Com '90 Hospitality will be open at Noon on Friday at the "Waters Edge" Lobby bar of the Sheraton CenterPark Hotel, adjacent to the Convention Center. At 7 PM Ham-Com will provide a giant round of roast beef and we'll draw the Preregistration Prize at 7:30. The "Lower Side Band", including our own K5VOU, will provide the genuine Texas country music. Be sure to come early enough to join in this great party.

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There will be testing for all levels from Novice to Extra Class. On Saturday there are sessions at 9 AM, 1 PM and 6 PM. The Sunday session begins at 9 AM. Tests are on a first-come basis. Bring your original license, a copy of your license which we may keep, and photo ID. The testing fee is $4.95.

Giant Indoor Flea Market

The giant indoor, air-conditioned flea market boasts 444 6-ft. tables with electricity available if desired. Tables are $15 each for the first three, $25.00 each for all over three. You will find everything you ever wanted from ham gear to computers, and much more. Don't wait to get your table, it's always a sellout!

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You will find 100 commercial exhibit booths at Ham-Com. A virtual who's who of amateur radio manufacturers. In addition you will find many national and regional dealers stocked and ready to sell you whatever you want. Bring money!

Ham-Com '90 Convention Banquet

Plan to attend the Ham-Com '90 Convention banquet at 7 PM on Saturday, June 9th. It's great food, a good time and a very interesting AMSAT program by Rich Ensign, N8IWJ, AMSAT Science Education Advisor. In addition, we will present the North Texas Section's Amateur Of The Year Award. Tickets are $16.00 each.

QCWA Continental Breakfast

Sunday morning, June 10th, at 7:30 AM, the Dallas and Fort Worth Chapters of the QCWA will host the QCWA Breakfast and Hospitality gathering at the Sheraton Hotel. Breakfast tickets are $3.50 each, but you need not purchase breakfast to attend the gathering.

Convention Schedule

Friday, June 8th, 1990:

12 Noon	Exhibit Set-Up, Ham-Com '90 Hospitality Open at Sheraton
2 PM	Ham-Com Registration Open
	Flea Market Open for Set-Up
7 PM	Hospitality & Entertainment at Sheraton.
9 PM	Set-Ups Closed

Saturday, June 9th, 1990:

6 AM	Ham-Com Registration Open
	Flea Market Open for Set-Up
7 AM	Flea Market Open
9 AM	Commercial Exhibits Open
11 AM	DX Lunch & Forum
2 PM	Xmtr Hunt
5 PM	Exhibits & Flea Market Closed
7 PM	Banquet at Sheraton

Sunday, June 10th, 1990:

7 AM	Flea Mkt./Registration Open
7:30 AM	QCWA Breakfast
9 AM	Commercial Exhibits Open
2 PM	Exhibits & Flea Market Closed
	Skywarn School

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Radisson Suite	$79	(817) 640-0440
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_______ Additional Tables, over 3, $25.00 ea
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_______ DX Lunch & Program, $10.00 ea
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THE GÖTTINGEN HEART ANTENNA

Uniquely shaped antenna gives good performance

By Walter J. Schulz, K3OQF, 15225 Wayside Road, Philadelphia, Pennsylvania 19116

One of the most interesting places in Germany is the valley where the Leine river wanders between the Harz mountains and the university city of Göttingen. To academicians, Göttingen is renowned as the city where 30 Nobel Prize winners have resided. It's also famous for the Max Planck Institutes located in the city and the countryside nearby. These 52 institutes are descendants of the Kaiser Wilhelm Institute organization which was devoted to the study of interesting scientific problems.

Gerd E.A. Meier, DJ7FY, and Rudolf J. Dvorak, DK4AP, of the Max Planck-Institut for Flow Research in Göttingen recently developed a new antenna. It is known throughout Germany as the Göttingen Heart (Herz) antenna, named for the city where it was developed and its heart shape.*

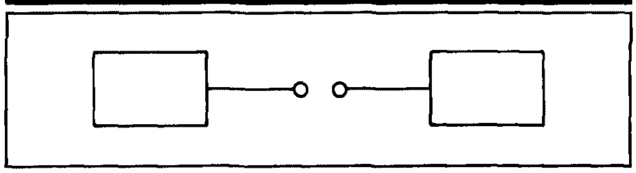
The antenna is unique. It is broadbanded over a more than 10:1 frequency range, and at the same time gives directive gain that increases with frequency. The antenna handles large amounts of RF power without corona discharge.

The Heart antenna doesn't need a large area to be broadbanded, as does the rhombic antenna, nor does it need the active dipoles that form a radiating cell of a log periodic array. The whole antenna is active on all excited frequencies and is small enough to be rotated on a single mast (see **Photo A**).

Dr. Meier was intrigued with professor Heinrich Hertz's original spark gap apparatus. Meier discovered through experimentation that two similar plates, when excited with RF energy, displayed not only directivity but some broadband behavior as well (see **Figure 1**). This led to the birth of the Heart antenna. Additional experiments by Dr. Meier and Mr. Dvorak showed that directivity was enhanced by

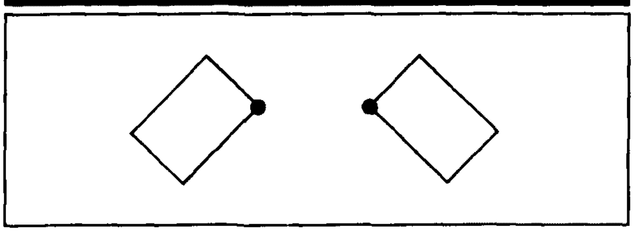
* The antenna is protected by EPü Patent number 0124559. An meldetag 15.11.83. Priority 15 Nov 82. Granted 4.03.87 Patent Blatt 87/10

FIGURE 1



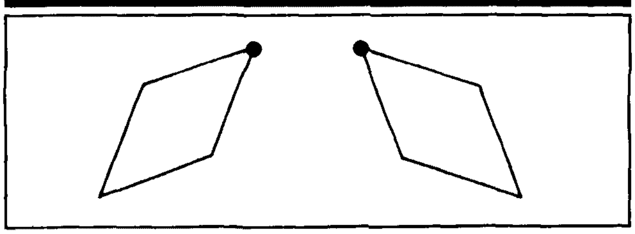
Spark gap with capacitance plates.

FIGURE 2



V configuration.

FIGURE 3

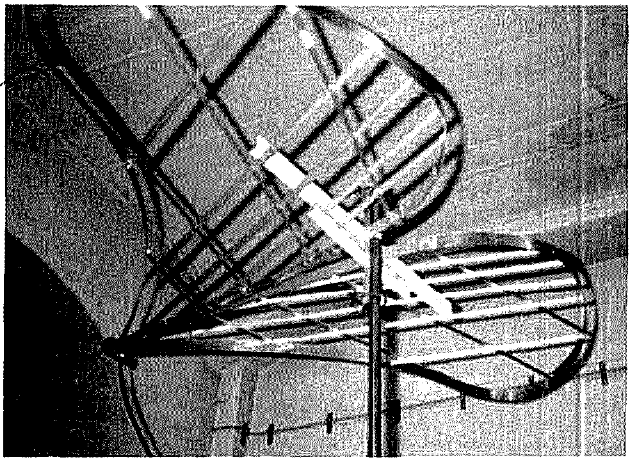


Rhombus plates in V configuration.

optimizing the V configuration of the plates, as shown in Figure 2. They also found that by changing from rectangular-shaped plates to rhombus-shaped ones (Figure 3), the antenna became broadbanded as the input impedance flattened out.

It was at this point that Mr. Dvorak found it necessary to "cut and try" many profiles. He noticed that as the curves of the antenna shape became smoother, there were less variations in the feedpoint impedance, and the VSWR ratio remained below 2:1. Moreover, as the frequency increased, so did the forward gain. At the same time, the antenna had approximately 16-dB loss off the back. The antenna's SWR was flat over the 10:1 frequency range and directive.

PHOTO A

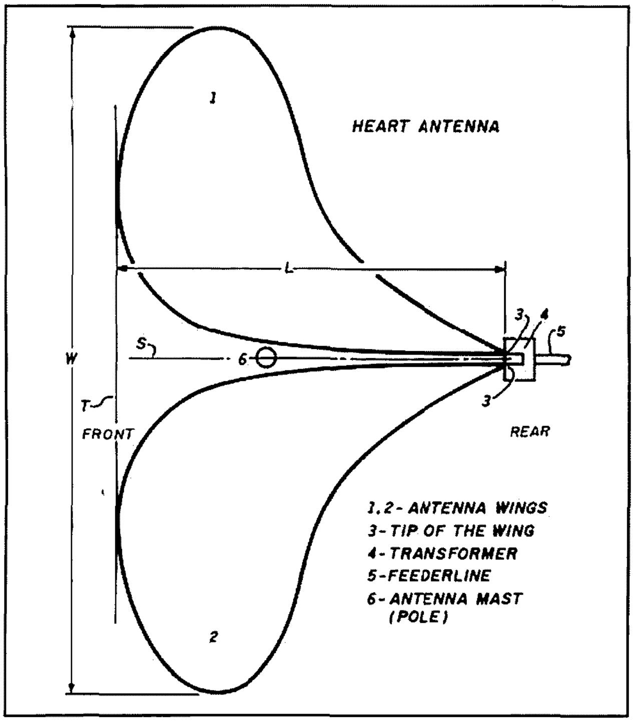


Active on all excited frequencies, the antenna is small enough to be rotated on a single mast.

The antenna wing profile

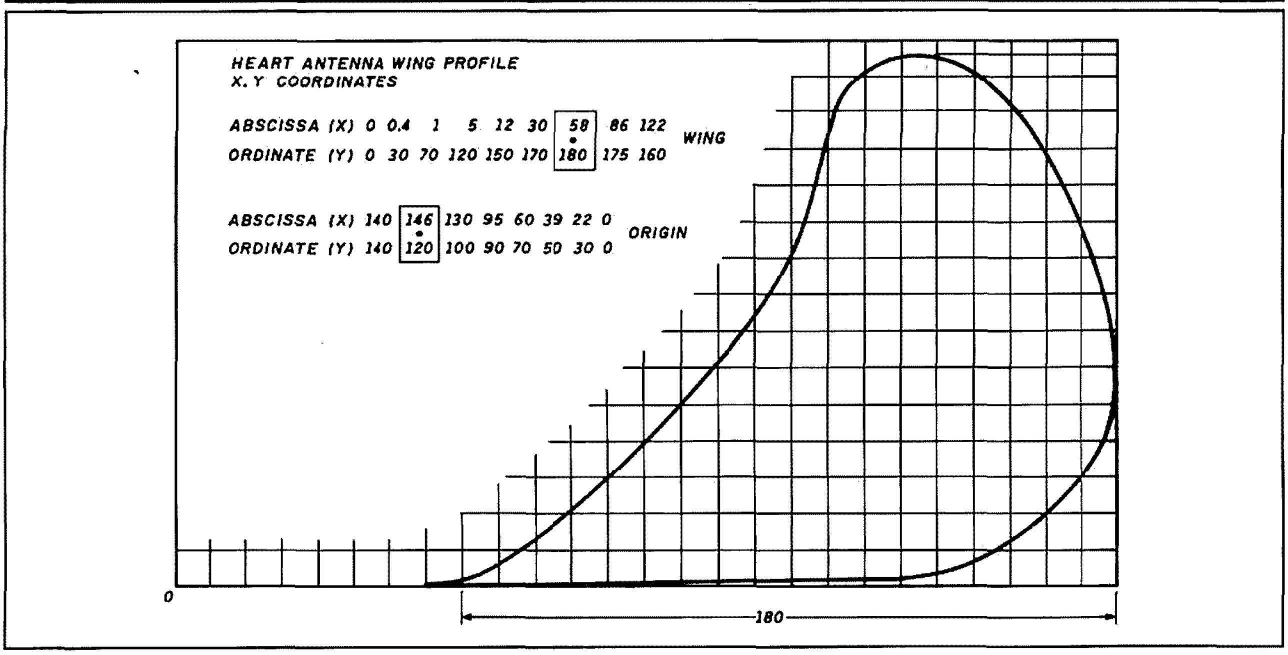
Figures 4A and B give the physical dimensions of the Heart antenna. Because this antenna is symmetrical, only one profile, or Heart wing, is shown in Figure 4B — a grid pattern.

FIGURE 4A



Symmetrical Heart antenna (showing both wing profiles).

FIGURE 4B



Heart antenna profile with X, Y coordinates.

You'll find that it's simplest to lay out a profile using graph paper. The grids on the paper make it easy to draw the curve shown in **Figure 4B**. Once you've made your sketch, you can transfer the profile to your building material.

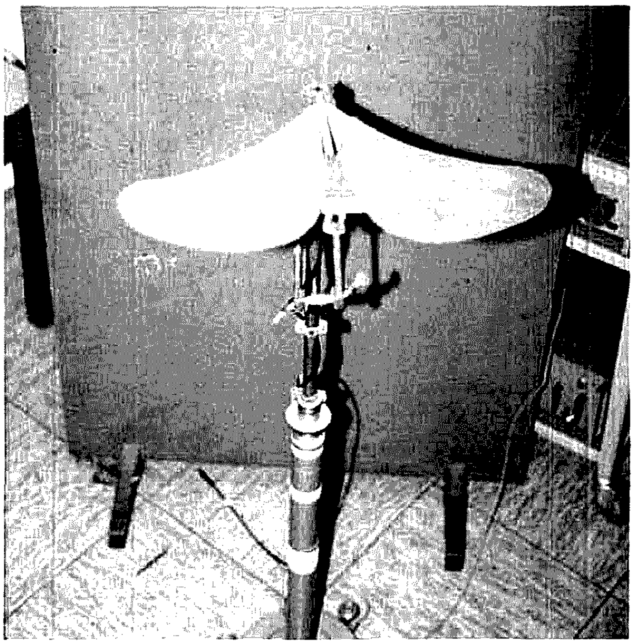
Always cut the profile to the lowest frequency of operation. The length (L) shown in **Figure 4B** is one-third of the longest desired operating wavelength, so you'll find the following relationship useful:

$$L = 1/3 \ maximum \ \lambda$$

Construction materials

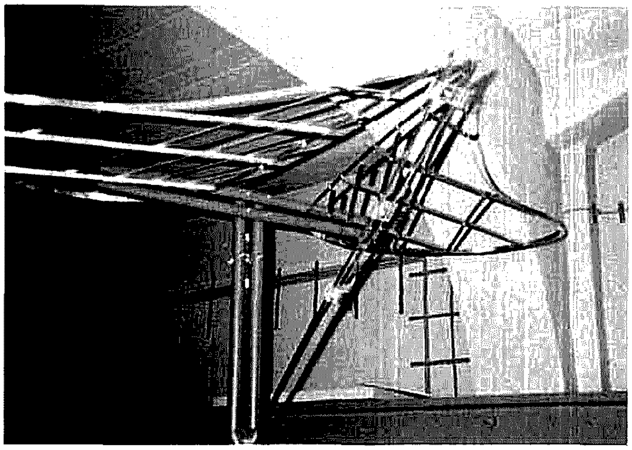
You can use common materials for antenna construction; they aren't critical electrically. This antenna has been built from aluminum foil, potato chip bag foil, chicken wire, wire mesh, foil-backed insulating foam sheathing, and aluminum angle stock (see Photos B and C). It is important to note that different material thicknesses and apertures will place certain restrictions on antenna performance, depending on the

PHOTO B



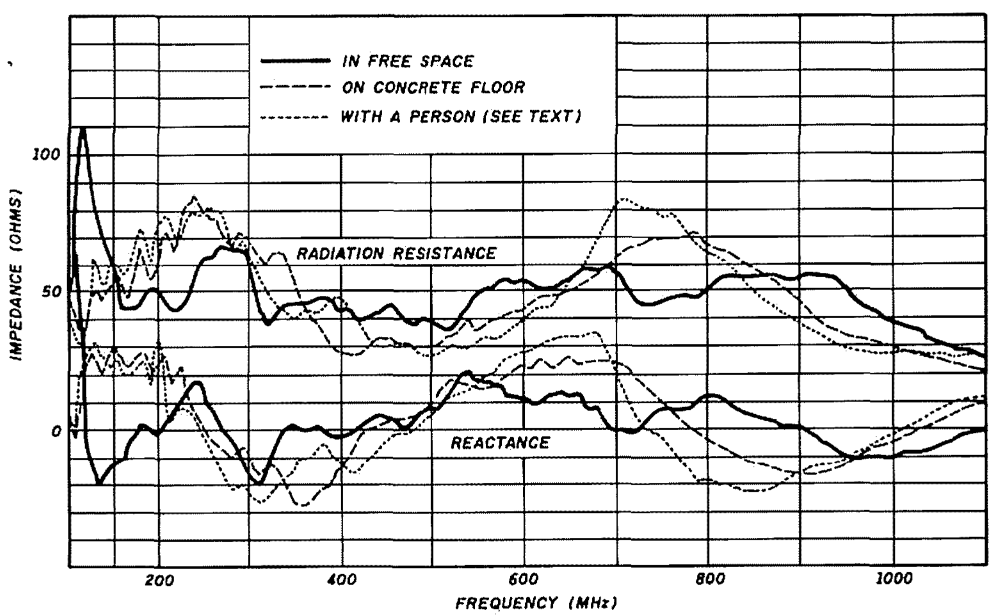
Common materials can be used to build the antenna. Here you see the Heart antenna made of aluminum sheet stock.

PHOTO C



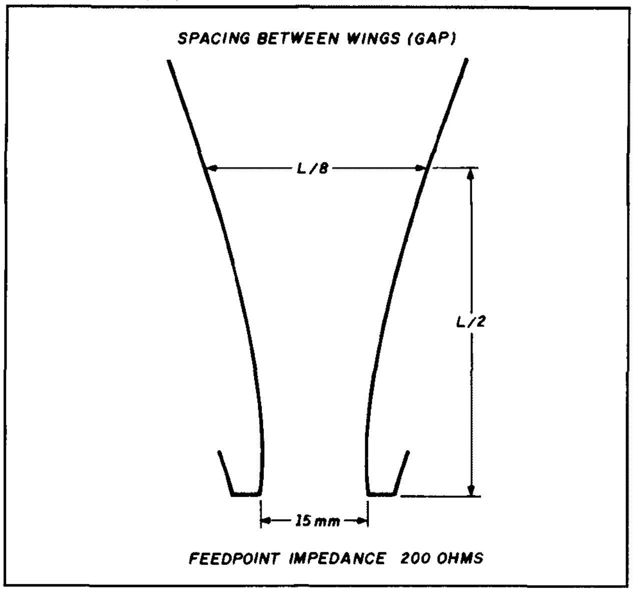
Another version of the Heart antenna constructed from aluminum strips.

FIGURE 5



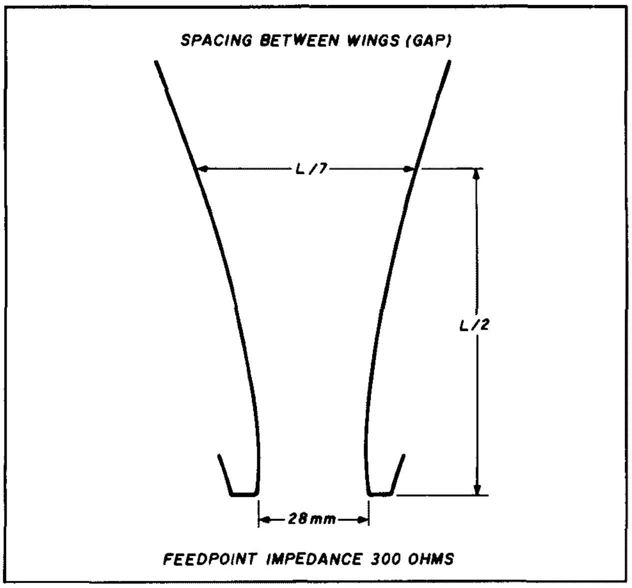
Impedance versus frequency of a symmetrical Heart antenna. (Antenna is made from sheet aluminum 1.5 mm thick.)

FIGURE 6



Profile wing spacing for 200-ohm feedpoint impedance.

FIGURE 7



Profile wing spacing for 300-ohm feedpoint impedance.

frequency and desired impedance at the feedpoint.

It's best to match the antenna with a step-down balun or an exponential line. An impedance transformation from 300 to 75 ohms, or 200 to 50 ohms, usually gives the best results and yields the highest efficiency. **Figure 5** shows feedpoint impedance versus frequency.

The spacing (gap) between the wings and the wing thickness determines the feedpoint impedance. A feedpoint impedance of 200 to 300 ohms is best. You can obtain a 75-ohm impedance by using close gap spacing between the wings, but this becomes very critical. The small distances involved are also affected by the wing thickness. **Figures 6** and 7 give the wing spacing for 200 and 300-ohm feedpoint impedances. These spacings are most efficient when the wings are made of thin materials — like foils less than 0.3 mm thick.

The gap must be larger for thicker materials or mesh wire with reinforced rims. Larger gap spacing — especially at half wavelengths — gives better gain at lower frequencies, but spoils impedance matching. The antenna impedance doesn't change when in close proximity to other objects. To prove this to yourself, you can place a foil Heart antenna on the floor, have someone stand on the antenna wings, and note the results.

Verification of results

Dvorak verified gain and impedance measurements with computer-controlled measuring equipment. He and Dr. Meier also submitted the antenna system to Gunter Schwarzbeck, DL1BU, the technical referent for the German Amateur Radio Club (DARC). Schwarzbeck, famous for his many critiques on radio antennas and measurements among European radio Amateurs, performed an impartial review, measurements, and a critical analysis. The Heart was tested thoroughly on a commerical antenna range. I have used the antenna here in the States and find it to be one of those antennas that wants to work and gives good results due to its simple construction.

In closing, I wish to thank Dr. Meier and Mr. Dvorak for making information about the Heart antenna available. You can address inquiries about the Heart antenna system to: Meier Messtechnik, Am Menzelberg 6, D-3400 Göttingen, Federal Republic of Germany.

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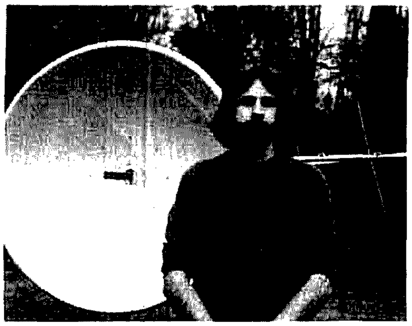
2. Gunter Schwarzbeck, "Herz-Antenne," *CQ-DL Magazine*, February 1988, DARC.
3. "Herz ist Trumpf," *Funk Magazine*, February 1985, pages 40-42.
4. Personal interview and correspondence with the inventors.



By Bob Atkins, KA1GT

Feedback

Although you're reading this column in late April or early May 1990, I'm writing it during the first week of February. Your letters are just starting to arrive and I am pleasantly surprised by their numbers and their overall message of support for the column. Some letters simply voiced their approval of a new column, while others contained pages of suggestions for topics. Please be assured that I read all the mail you send and will take your comments into account when planning the subjects to be presented in future columns.

Some readers (I know of at least one!) will be glad to learn that this is the last column devoted to optical communication. Though optical systems aren't strictly "microwave" frequencies, I decided to cover them for a couple of reasons. First, contacts at optical frequencies do qualify for points in a number of ARRL-sponsored contests and therefore have some legitimacy in Amateur Radio. Second, there are quite a few Amateurs experimenting with optical communication (indeed a ham radio talkback link is almost essential in setting up a DX optical link). Third, I felt there was a need to make available information which I haven't seen presented in the literature generally available to radio Amateurs. I may come back to the subject from time to time, but next month the column will move down in frequency by several orders of magnitude and return to the traditional microwave bands!

One concern expressed in a number of letters is the problem of finding local help for those trying to get started on the microwave bands. I might be able to be of service if the secretaries of VHF/UHF/Microwave clubs and organizations send me copies of their membership lists (paper copies or MS-DOS readable disks — any size, any capacity, ASCII, dBASE, most word processing formats). I could then try to direct inquiries to the closest active microwave station in the area.

Thanks to the organizations who sent along copies of their newsletters. A lot of information appears in these newsletters which is seen only by a small number of hams. I'll try to extract short items (with full credit given, of course) for this column, or in the case of longer technical items, I'll print short abstracts and direct readers to the original articles. If you send in a newsletter, please include instructions on subscribing so I can pass the information along.

A few items of interest from this month's newsletters include:

The 24th Annual Central States VHF Conference will be held in Wichita, Kansas at the Marriott Hotel, July 26 through 29, 1990. In addition to the technical program, there will be antenna and noise figure measurement sessions. KFØM is organizing the technical program and any contributions to the proceedings (to be published by the ARRL) should be submitted to him by May 15th (as ASCII files on IBM compatible 5-1/4" diskettes). KD5RO/2 is collecting prizes for the conference. (MWVHFR)

After a successful meeting in the Dallas/Fort Worth area in 1989, the 1990 Microwave Update meeting moves back to Colorado and will take place somewhere in the Denver area under the care of WØPW. I'll have more on dates and location when they are announced. (MWVHFR, NTMS)

OE9PMJ published details of a 76-GHz station in *DUBUS*, April 1989. The rig starts with an 81-MHz crystal oscillator, multiplied and amplified to give 1 watt at 978 MHz. This is then multiplied by 16 in a diode multiplier (BXY 21) and amplified (MGF1302) to give 10 mW at 15.6 GHz. This is fed to a waveguide 5x multiplier/mixer/ filter. Output power is a few microwatts and noise figure is around 40 dB. FM modulation of the crystal oscillator is used and operation is in the Gunnplexer mode (two stations use offset frequencies equal to the receive IF). With 35-cm dishes, DX is about 3 km so far. (NTMS)

AA5C reports using a 45-foot run of 3/4-inch copper tubing as circular waveguide at 10 GHz with a loss of only 1 dB. WR-75 transitions to rectangular guide were used at the ends. By flattening the ends of the tube, it was possible to insert the WR-75 rectangular guide and solder the two together. Total cost was around $32. (NTMS)

N3CX has designed a 903-MHz DL6WU-type Yagi. Construction details are given. (PACKRATS via NTMS)

Credits:

NTMS — North Texas Microwave Society "Feedpoint" (January/February 1990). NTMS membership is $12 per year. Contact Wes Atchison, WA5TKU, Rt. 4, Box 565, Sanger, Texas 76266.

MWVHFR — Midwest VHF Report (November 1989). Subscription rate is $7.50/6 months, $15/year. Contact Roger Cox, WBØDGF, 3451 Dudley Street, Lincon, Nebraska 68503-2034.

PACKRATS — Mt. Airy VHF Radio Club "Cheese Bits" (September 1989). Subscription rate is $5 per year. Contact Doc Cutler, K3GAS, 7815 New Second Street, Elkins Park, Pennsylvania 19117.

Other news

The RSGB seems to have beaten the ARRL to the punch with the publication of the first part of their *Microwave Handbook*. I haven't seen a copy yet (I have my order placed!), but I'll review it in this column as soon as I receive it. By the time you read this it should be available from the HAM RADIO Bookstore.* As of early February 1990, the first part of the long-awaited ARRL *Microwave Handbook* is still in the preparation stages, but rumor has it that it's almost ready and "it won't be long now" before it's in print.

Atmospheric optical propagation

In this, the final part of a series of columns on laser communication, I'll take a look at the atmospheric propagation of light. While there are many propagation modes for RF energy — like E skip, meteor scatter, line of sight, troposcatter, and ducting — there's

*Available from the HAM RADIO Bookstore for $35 plus $3.75 shipping and handling

only one mode of propagation for light that's of practical importance in optical communication. This mode is line of sight. While propagation via tropospheric scattering is possible, the equipment needed (several watts of laser power and lenses or mirrors many feet in diameter) is beyond the reach of Amateurs. The only non line-of-sight mode of optical propagation which could be within the reach of Amateurs is a form of ducting known as a mirage. It occurs when layers of air of very different refractive indexes cause light to bend around the curvature of the earth. If you can recall the number of mirages you have seen, you'll get some idea of the low probability of using this propagation mode for laser communication!

Restricting the discussion to line-of-sight propagation, you must first ask what effects the atmosphere has on a laser beam. For example, does the atmosphere absorb light? At visible wavelengths, there's no significant absorption of light. This isn't true in the ultraviolet and infrared regions, but we're only concerned here with visible light. Another effect of the atmosphere on light propagation is due to turbulence. Stars "twinkle" not because they vary intrinsically in brightness, but because their light undergoes rapid intensity and positional changes as it passes through the unhomogeneous atmosphere. Similar effects may be observed on a laser beam passing through the atmosphere. However, for the simple communications system described here, they will not affect the received signal as long as they aren't so severe as to result in a laser beam deflection great enough to cause it to miss the receiver for a significant fraction of time. This is very unlikely to occur and can be disregarded for all practical purposes.

Scattering is the only atmospheric effect on light propagation of significant importance to a simple optical communications system. Light which is scattered by the atmosphere is prevented from reaching the receiver and results in received signal attenuation. There are two causes of atmospheric scattering. First, there's scattering by the molecules of the air itself. This is fairly constant and can be described by the process of Rayleigh scattering (scattering by elements much smaller than the wavelength of the radiation being scattered). Second, there's scattering by particulates in the atmosphere — water droplets and dust particles, for example. This type of scattering is highly variable and is the cause of smog and haze, with which we are all familiar. It can be described by the process of Mie scattering (scattering by elements similar in size to the wavelength of the radiation being scattered). Both types of scattering result in higher losses for shorter wavelengths; that is, blue light is scattered more than red light. This is why the sun (and often the moon) appears very red when rising or setting. The red light penetrates the atmosphere with less scattering losses than the blue light. Similarly, the sky is blue because we are seeing the preferential scattering by the atmosphere of blue light.

In addition to scattering, the other significant process which gives rise to signal loss on a line-of-sight path is due to the geometric spreading of the laser beam. If you know the path length, the divergence of the beam, and the size of the receiving optics, you can calculate how much of the original laser power will be collected by the receiver using simple geometry that I'll describe later. If a beam expander is used, then the transmissivity of the optics involved must be taken into account, but this is usually just a small correction.

If you're interested in the details of the calculations involved, read the appendix to this month's column. The math is simple geometry and algebra, but I won't inflict it on those it might scare away!

The most important variable in transmission is the clarity of the atmosphere. In my area, New Jersey, visibility of 20 miles (as noted in local weather reports) is not uncommon. I have heard reports of 50+ mile visibility while in Maine, and on a recent trip to California I heard a weather report claiming 100-mile visibility in the Sacramento Valley.

Using the formulas given in the appendix or the graphs derived from them to calculate the range of a typical system, you should come up with the following:

Given: Laser power = 1 mW
Laser beam divergence = 1 mR
Laser wavelength = 632.8 nm (He-Ne)
Diameter of receiving optics = 8 inches
Transmissivity of receiving optics = 0.9
Detector sensitivity = 1×10^{-10} watts
Visual range = 32 km (20 miles)

The DX range at a 0-dB signal to noise ratio is 49 km. This reduces to 45 km for a 3-dB ratio and to 40 km for a 6-dB ratio. Doubling either the laser power, receiving optics area, or detector sensitivity will only increase the 0-dB S/N range to about 54 km. Multiplying any one of these same factors by 5 yields an increase in range to 61 km. However, an improvement of only 25 percent in visual range (to 40 km) yields a DX range of 58 km with the original equipment parameters. Waiting for a slightly clearer day can result in the same difference in potential range as will a large change in equipment, and small increases in distance can make a big change in signal strength. Note that using blue light (440 nm) in place of red indicates a reduction in range of around 40 percent under these conditions.

I hope that this analysis gives you some idea of the capabilities of modest laser communications systems and the factors important to performance. The numbers calculated should be used only as a guide, because the calculation of scattering losses on the basis of "visual range" is only an approximation.

Appendix

To calculate the strength of a laser signal over a given path, you need to know the following:

1—The laser power
2—The beam divergence
3—The transmissivity of any beam expander
4—The path length
5—The transmissivity of the atmosphere
6—The area of the receiving optics
7—The transmissivity of the receiving optics

All of these factors are known (or can be reasonably estimated) with the exception of the atmospheric transmissivity loss factor. This factor can be calculated using the formula:

$$\textit{atmospheric transmissivity} = e^{-(M+R)*D}$$

where M is the Mie scattering coefficient (km^{-1}),
R is the Rayleigh scattering coefficient (km^{-1}),
D is the path length (km), and
(M+R) is the atmospheric attenuation coefficient.

You can estimate the scattering

coefficients using empirically derived relationships. The Rayleigh scattering coefficient is given by the expression:

$$R = 0.00833*(\lambda/600)^{-4.08}$$

where λ is the wavelength of the laser light (nm).

The Mie scattering coefficient is given by the expression:

$$M = (3.91/V)*(\lambda/550)^{(-0.585)*(V^{0.3})}$$

where V = the visual range (km).

Visual range is a measure of how far you can see, based on the apparent contrast observed in a standard target at a given distance. It's often given in weather reports and aviation weather forecasts. It may be estimated on the basis of the following qualitative descriptions.

Conditions	Visual range (km)	Atmospheric attenuation coefficient (km^{-1} at 632 nm)
Extremely clear	100	0.03
Very clear	40	0.08
Clear	20	0.16
Light haze	7	0.49
Haze	2.5	1.40

This information is also depicted graphically in **Figure 1**.

Referring to **Figure 2**, you have a laser of power P watts and beam divergence θ radians illuminating an area of radius r km at a distance of D km.

For small angles,

$$Tan[\theta(degrees)] = \theta(radians)$$

The radius, r, of the illuminated circle at a distance, D, is given by the relationships:

$$r = D\ Tan(\theta/2)\ [\theta \text{ in degrees}]$$
$$r = D\ \theta/2\ [\theta \text{ in radians}]$$

The area illuminated will therefore be given by the relationship:

$$area = \pi r^2 = (\pi D^2\theta^2)/4$$

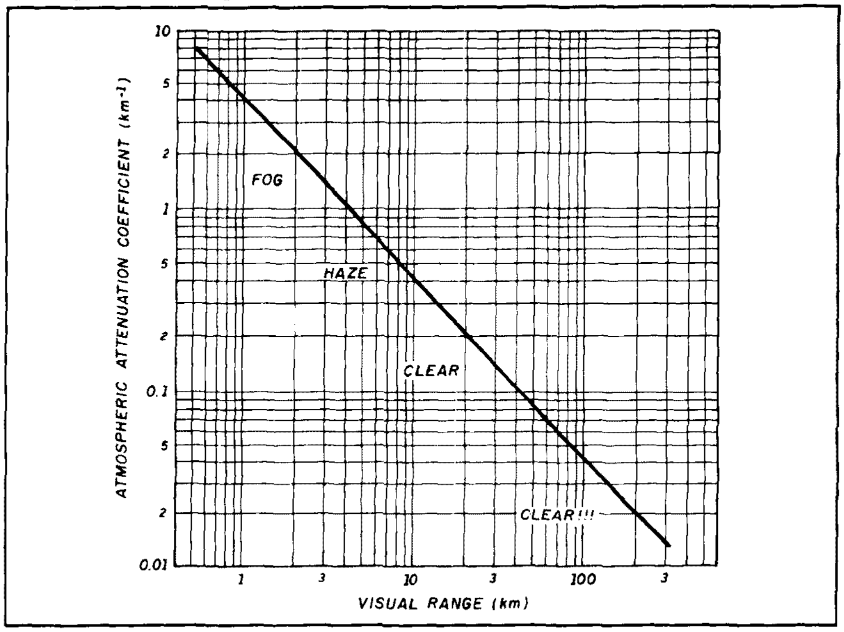
The power density at the target is given by power/area and is therefore (assuming uniform illumination):

$$power\ density = 4P/\pi D^2\theta^2$$

(In fact the power density will be somewhat higher if the laser is aimed accurately, because the illumination isn't uniform across the beam but peaks in the center. Thus, this is a conservative estimate.)

For a detector capturing light from A (area of lens or mirror), the power received will be given by:

FIGURE 1



Relationship between visual range and atmospheric attenuation coefficient. Values are approximate. They will be about 10 percent lower at the red end of the spectrum and about 20 percent higher at the blue end.

$$power\ received = 4PA/\pi D^2\theta^2$$

And, if To is the transmissivity of the optics and Ta is the transmissivity of the atmosphere, then the power incident on the detector will be given by:

$$power\ at\ detector = 4PAToTa/\pi D^2\theta^2$$

If the detector can detect a minimum signal power Dp, then at the maximum communication distance D:

$$Dp = (4PAToTa)/(\pi D^2\theta^2)$$

Rearranging this, you have:

$$D^2 = (4PAToTa)/(\pi Dp\theta^2)$$

The only unknowns in this equation are D, the maximum communications distance, and Ta, the transmissivity of the atmosphere. Ta can be determined as described above, and so D can be calculated.

Ta is given by the expression:

$$Ta = e^{-(M+R)D}$$

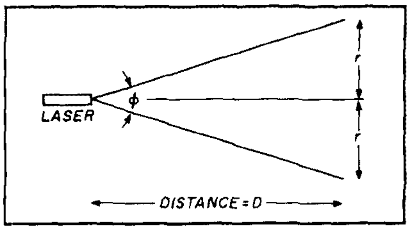
where e = 2.71828, M is the Mie coefficient, R is the Rayleigh coefficient, and (M+R) is the combined atmospheric attenuation coefficient.

Thus, the final expression for calculating maximum distance is:

$$\frac{D^2}{e^{-(M+R)\,D}} = \frac{4PATo}{\pi Dp\ \theta^2}$$

This isn't a trivial equation to solve.

FIGURE 2

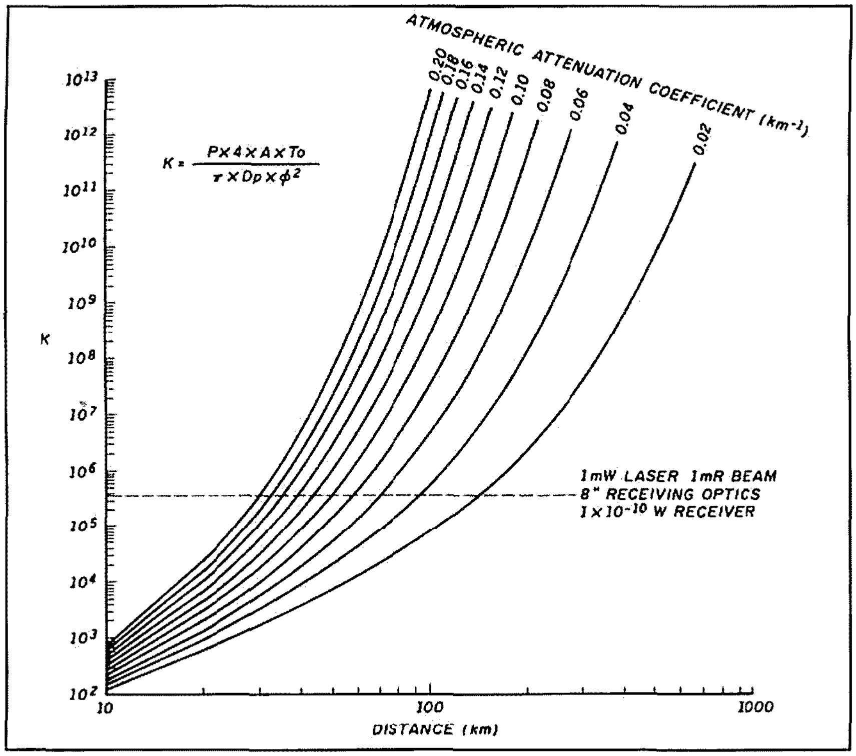


A graphic depiction of the laser beam divergence illuminating an area.

It can be solved graphically using **Figure 3**, where the right-hand side of the equation is plotted against DX potential for a number of different scattering conditions. Note that units must be consistent. Here's an example: for distance in km, the area of the receiving optics (A) must be expressed in square km, the atmospheric attenuation factor (M+R) must be in units of km^{-1}, and all power and sensitivity factors must be expressed in the same units (watts). Perhaps the easiest solution for those with access to a computer is to use a trial and error method. Because all quantities except D are known, substitute increasing values of D into the equation until both sides are equal. A TURBO BASIC listing* is available

*Send an SASE with $.25 postage to *Ham Radio* Magazine for program listing

FIGURE 3



Relationship between K (as defined above) and maximum communication range for an optical communications system for a number of different atmospheric conditions (as given in *Figure 1*).

which uses this formula to calculate DX range (at 0-dB signal to noise ratio) for a given atmospheric communication system. The program should also run under QuickBASIC, but to run under more traditional versions of BASIC (like GW-BASIC) some modifications, like the addition of line numbers, the replacement of labels with line numbers in the GOTO statements, and the REM keyword before the remarks, will be required. A compiled version of this program which will run on any IBM PC compatible machine is available from the author for $5 to cover materials and shipping. hr

ROTATABLE TRAP DIPOLE
FOR 12 AND 17 METERS

By Gary Nichols, KD9SV, 4100 Fahlsing Road, Woodburn, Indiana 46797

Like many Amateurs, I tried loading every antenna I had on 12 and 17 meters with an antenna tuner when we were given operating privileges for the new WARC bands. I had limited success. My small (33 foot) off-center loaded dipole for 40 meters was about the best I had. However, I've never been fond of hiding a mismatch with an antenna tuner.

I decided to redo my antenna system and convert the 40-meter rotatable dipole into a trap antenna for 12 and 17 meters. I also decided to shunt feed the tower as a 5/8-wavelength vertical on 40 meters; this worked out better than the dipole.

Trap description

Design guidelines suggested that 100 to 300 ohms reactance is recommended for trap design. The capacitance of the tubular traps is about 37 pF, which is very near 200 ohms at 24.95 MHz. Fourteen turns of no. 12 solid housewire will resonate with the tubular capacitor on 12 meters (see **Figure 1**).

The trap is built with two telescoping sections of aluminum. A piece of polybutylene water pipe is placed between them for the capacitor dielectric. This makes a self-supporting, adjustable, easily constructed trap. It's similar to the gamma capacitors used on some beam antennas. The coil is wound over part of the plastic pipe.

Assembly

Cut all aluminum pieces in half. Use a hacksaw to cut two lengthwise slits, about 1 inch long, in one end only of the 1-inch and 7/8-inch tubing. These are for the hose clamps. Make four slits in the 3/4 and 1/2-inch tubing on one end only.

Cut two 8-3/4 inch pieces of the plastic tubing and "hammer" them onto each piece of 1/2-inch aluminum. Leave 1 inch of aluminum exposed at the slit end. You'll have to stand the tubing on end on top of something hard and pound the plastic on, because it fits very snugly.

Take a 3-foot length of no. 12 solid wire with black insulation and wind about 17 turns over a 1/2 or 9/16-inch diameter tube. Then screw the coil over the plastic and assemble as shown in **Figure 2**. Use 14 turns for the coil, and terminate the coil ends with a strip of tin-plated brass. Wrap the tin plate around the aluminum tubes and secure with a hose clamp. With the 1/2-inch aluminum tube engaged approximately 4-3/4 inches into the 3/4-inch tube, the trap should resonate near 25 MHz. Tune the trap by sliding it in or out slightly; set resonance at 24.9 MHz using a grid dip meter. Be sure to check your dipper with the station receiver for accuracy.

PARTS LIST

1-inch OD 0.058-wall aluminum, 6 feet
7/8-inch OD 0.058-wall aluminum, 8 feet
3/4-inch OD 0.049-wall aluminum, 8 feet
1/2-inch OD 0.058-wall aluminum, 17 inches
3/8-inch OD 0.058-wall aluminum, 6 feet

5/8 × 1/2-inch plastic water pipe, polybutylene, 2 pieces approximately 9 inches each

8 hose clamps: 4 each 1 inch, 2 each 3/4 inch, and 2 each 1/2 inch (all stainless preferred)

12 inches of PVC pipe approximately 1-inch ID, schedule 40 pipe

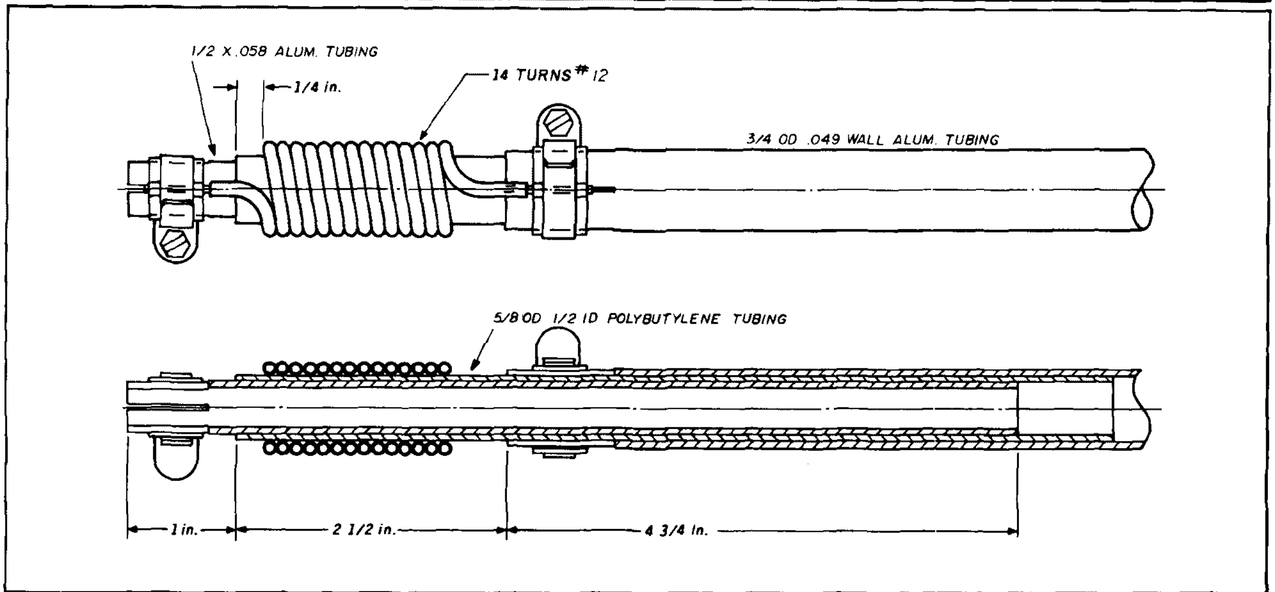
4 U bolts to fit over the PVC pipe

Angle or bar stock for antenna-to-mast mounting plate (I used 1/4 × 3-inch × 12-inch aluminum.)

0.015-inch thick tin-plated brass stock (4 pieces 1/2 × 2 inches) for terminating coil ends

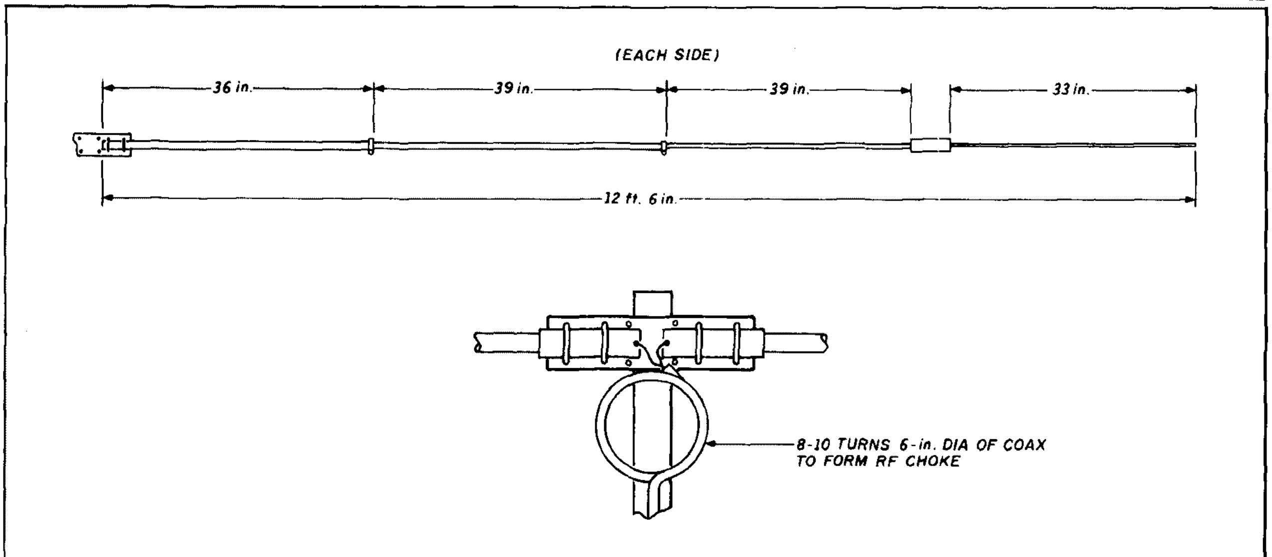
6 feet of no. 12 solid wire with TW or TWH insulation (black)

FIGURE 1



The 12-meter trap.

FIGURE 2



Assembly of 12/17-meter rotatable dipole.

Don't use white insulated wire; it's not very ultraviolet resistant. Be sure to use the tin-plated brass to terminate the coil ends. This will give you a lot of contact area and it won't corrode with the aluminum, as copper will. If you follow the drawing dimensions closely, almost no further adjustment will be necessary. I placed the trap and 4-foot section of 3/4-inch tubing on the top of a plastic wastebasket while checking with the grid dip meter.

Finish assembling the antenna as shown in **Figure 2** and test it with the antenna up about 20 feet or so, if possible. Input impedance on a properly tuned antenna should be approximately 60 ohms on 12 meters and about 65 ohms on 17 meters for a VSWR of about 1.3:1.

For best results feed the dipole with a good piece of coax, and use either a 1:1 balun or RF coax choke. Eight turns 6 inches in diameter are adequate.

This antenna has taken a kilowatt keydown with ease, and the trap capacitor has been hi-pot tested to 25 kV. Full legal power shouldn't be a problem.

I have completed antennas available, should you not wish to build your own. Write to me at the address at the beginning of this article for details.

THE HELICAL WHIP FOR RV AND MOBILE USE

By J.T. McCullough, WØBHG, 1401 North Jesse James Road, Excelsior Springs, Missouri 64024

When I retired I found I needed to reduce the complexity of my antennas for several bands, and simplify their mounting and connections. I needed to cut down on the time I spent setting up and disassembling antennas for recreational vehicle (RV) and motor home use.

Advantages

The antennas presented here will also work well for ordinary mobile operations. They can be adapted easily for home use in tight space situations in an attic, on a porch roof, or on a small lot. While it's quite likely that the helical vertical will never perform any better that a full quarter-wave vertical in the same situation, the difference can be rather negligible. Helicals can also be used back to back for a one-half wave vertical or horizontal dipole. They will work as radials for a ground plane antenna too.

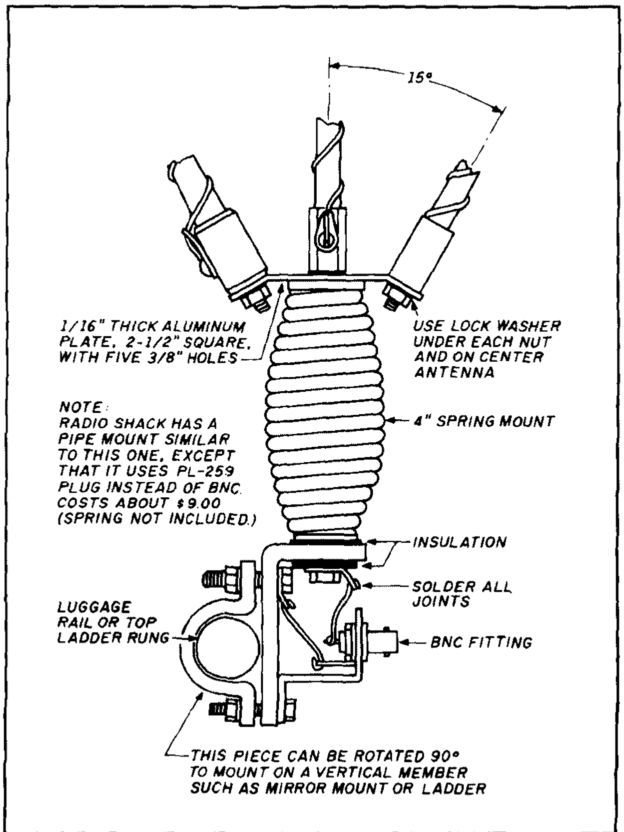
For temporary or permanent use on one WARC band, attach the helical vertically to the framework of a metal beam assembly near the feedpoint. Then connect the base directly to the coax feed. The framework and elements of the beam provide the ground plane. By mounting the helicals at a slight angle from each other, you could use two or three new bands in the same way.

It appears that at certain times of day, especially in the early evening, stronger DX signals may be obtained from 10, 15, and 20-meter helicals atop a metal-covered RV than from much more sophisticated antennas in the same location. This may be because the metal covering of an RV makes an almost perfect ground plane.

Development

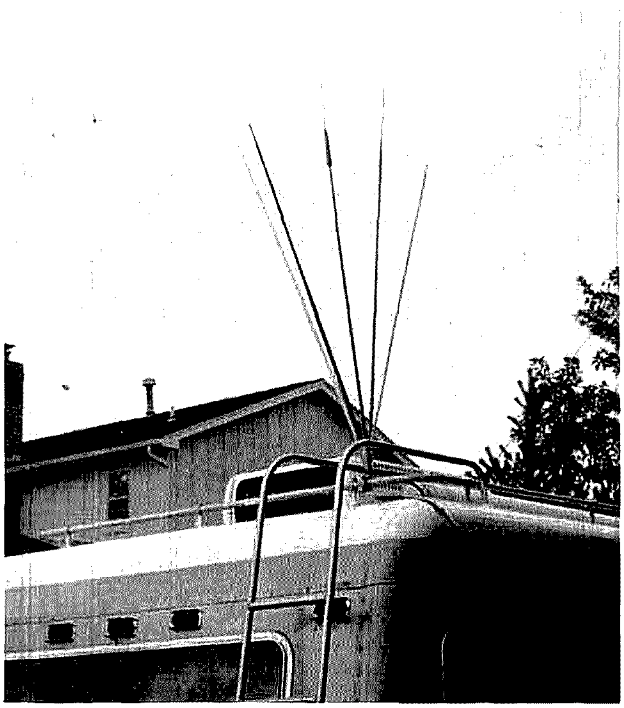
I started using helical whips with a large magnet mount about two years ago. My first was a CB antenna, modified

FIGURE 1



Motor home antenna mount.

PHOTO A



Five-band helical antenna on single spring mount.

for 20 meters. It was fantastic! I used it atop my pickup truck while traveling, and on my trailer or motor home at campsites. I worked a lot of DX, including all continents. (More about modifying the CB antenna later.)

I wanted to work five bands: 10, 12, 15, 17, and 20 meters. To do this I tied all five helicals together near the feedpoint. I started with a heavy 4-inch spring mount. Then I made a 2-1/4 inch square plate out of 1/16 inch thick aluminum and drilled five 3/8-inch holes — one in the middle and one in each corner. I bent each corner down about 15 degrees. I installed one antenna in the center hole to hold the plate in place on top of a spring mount. Next I put one antenna in each of the other four holes, using lock washers and 3/8-24 nuts. See Photo A and Figure 1 for details.

Performance comments

There appears to be no significant interaction between the five antennas, with the possible exception of the center unit. Each was developed individually using a method I'll describe later, so that VSWR was lowest in the center of the desired band. When attached to the common spring mount on top of the motor home, four of the antennas had low VSWR (less than 1.5:1) without retuning.

The center unit (20 meters) was affected slightly, and I needed to shorten the tip 1/2 inch to obtain the best VSWR for the phone band. The 20-meter antenna also had narrow bandwidth, although it still covered the phone band. For this reason, it might be better to put one of the narrowband antennas (11 or 17 meters) in the center.

While traveling, tilting the assembly about 45 degrees or more to the rear should help you avoid losing some of it to an underpass or bridge. You'll encounter another hazard in residential areas, where large trees often overhang the streets. Mobile operation is still quite feasible, even at a 45-degree tilt, if you mount the assembly on a rear luggage rail or ladder.

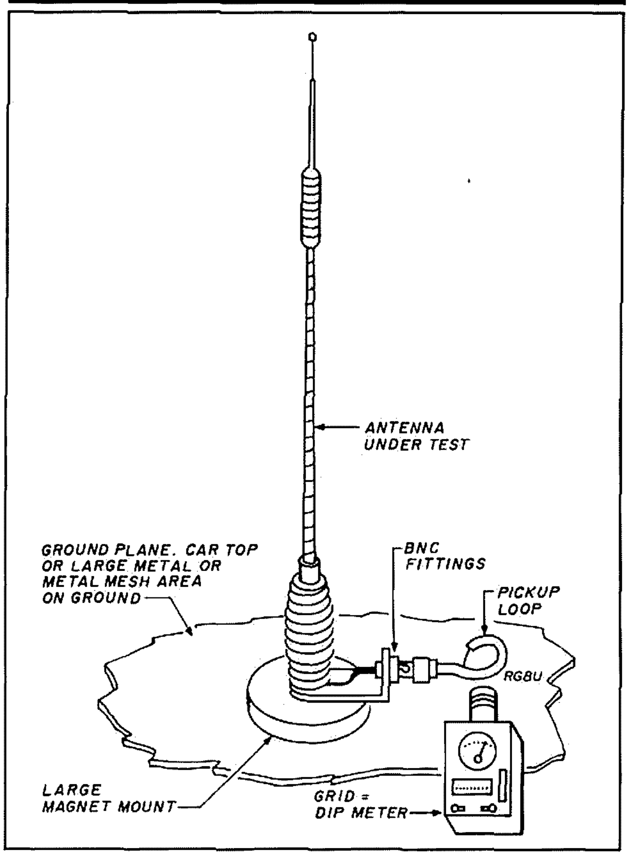
Testing procedure

I used the test setup in Figure 2 to check all the antennas during construction. If you don't plan to go into heavy production, the steel top of a car or pickup truck should work quite well as a base for the magnet mount. Take precautions to protect the top of the vehicle. A thin sheet of plastic should work, as capacitance between the magnet mount and car top will serve as an effective RF ground connection.

I consider a grid-dip meter a must for checking antenna resonant frequency, but an RX noise bridge may work if you are proficient in its use. Checking VSWR is useless for initial testing unless the resonant frequency happens to be in or very near the band in question. If you have a BNC fitting on the magnet mount, you can easily attach a small loop for grid-dip meter readings, exchanging it for coax for testing VSWR or on-the-air use.

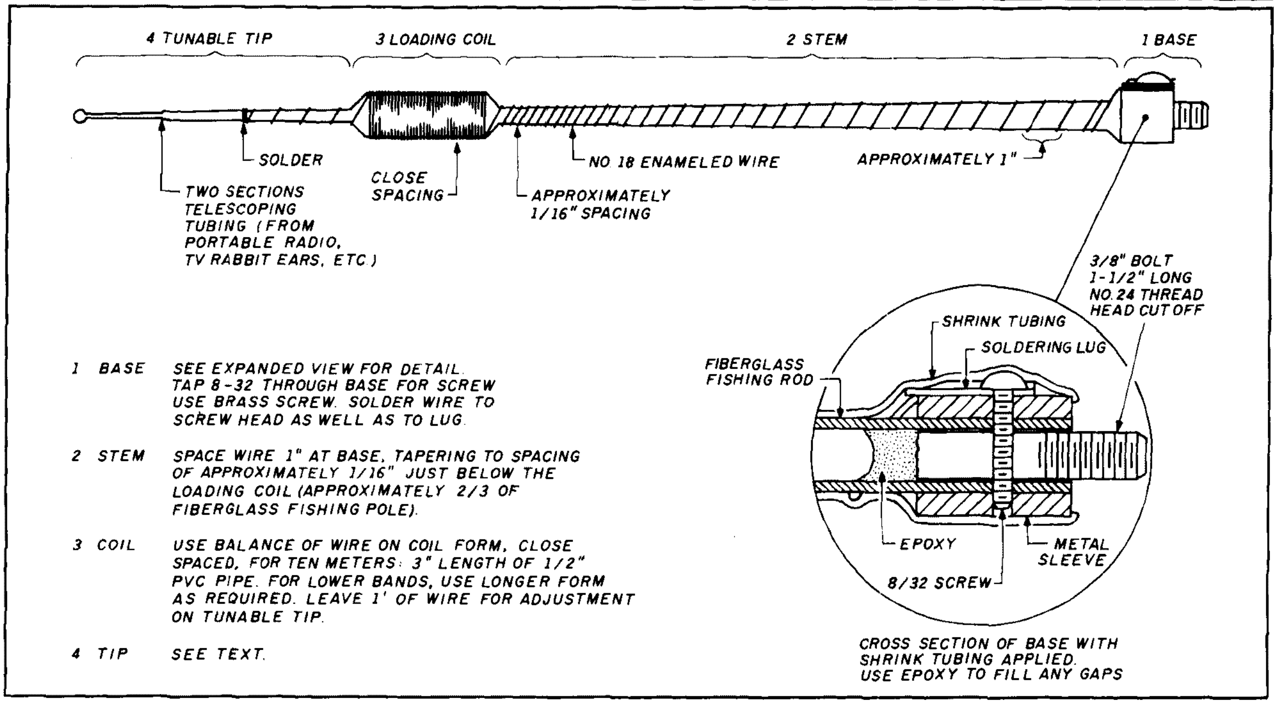
After you adjust an antenna roughly to the correct frequency with the grid-dip meter, make fine adjustments by connecting the antenna to a transceiver and finding the frequency of lowest VSWR with a VSWR meter. An adjustment of as little as 1/4 inch to the tip of the antenna may

FIGURE 2



Test setup.

FIGURE 3



Fiber glass helical antenna.

be necessary to bring the lowest VSWR point to the center of the desired band. It's usually sufficient to check the VSWR every 100 kHz to find the low point. A dual needle VSWR meter is a real time saver here.

It's better to start with an antenna that's too long; you can find the correct length with careful pruning. If you haven't found a low VSWR point after your first check with a grid-dip meter, recheck the meter's frequency. Don't be concerned with other higher frequency dips. Some may be harmonic dips and some may not. Depending on the length of the wire up to that point, the loading coil may act as a trap or a choke, creating a much higher frequency dip.

Construction of the modified CB antenna

The modified CB antenna requires the least work and is very efficient. I recommend using a top-loaded type with short tuning stubs at the very top, just above the loading coil. The one I used had about 3 feet of helical winding spaced evenly on constant diameter fiber glass rod, followed by a close-spaced loading coil several inches long, and topped by a metal tuning tip with a set screw and 2 or 3 inches of 1/8-inch steel rod. I replaced this rod with one of the same diameter about 24 inches long, and cut and adjusted it until it resonated at the proper point in the 20-meter band.

It's important not to extend this rod into the loading coil, as it will cause frequency change. If it is steel, it will likely heat up if much power is applied. Your local hardware store is a good source of brass brazing rod. While I didn't try this antenna on other bands, it should work well on 12, 15, or 17 meters if you use different lengths of preadjusted rod. For 10 meters, it will probably be necessary to remove a few turns from the loading coil.

Neighborhood garage sales and flea markets are good, inexpensive sources for CB antennas. Because most late model, top-loaded CB antennas don't have the tuning tip, you may need to improvise one (details on how to do this later).

Building fiber glass rod helicals

The light weight of fiber glass fishing rods makes them useful in helical antenna construction. Weight can be important when four or five antennas are placed on one spring mount.

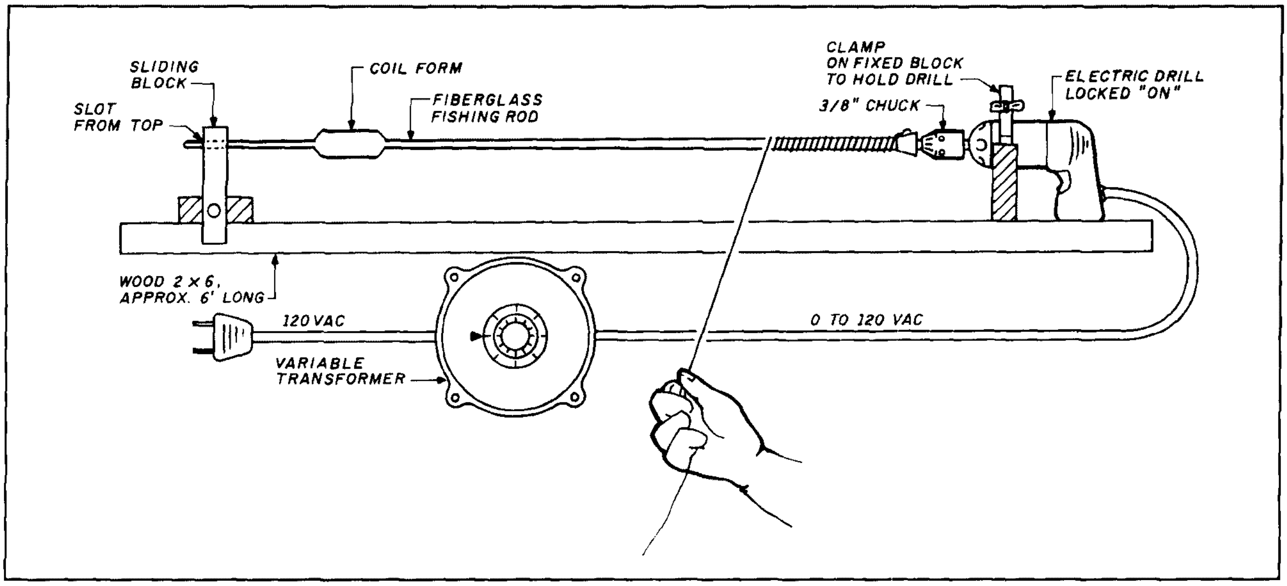
Choosing a fiber glass rod

You can find fiber glass rods at garage sales and flea markets. Poles with broken or missing guides and handles can often be bought for almost nothing. Look for hollow rods at least 4 or 5 feet long with a 3/8 to 1/2-inch diameter near the handle. You can use smaller, solid rods, but they require heavier, larger loading coils. Use one-piece rods with no metal ferrules or joints. It's possible to use larger rods with metal ferrules if you cut them out and rejoin the rod pieces with epoxy. One way to do this is to slide one end into the other, or use a tight-fitting piece from a solid rod as an internal splint. Refer to **Figure 3** for construction details.

Base

Remove all windings, wire, and thread, along with the tip, guides, and any foil or metal bands. Remove the entire handle, unless it appears that an inch or two would make a good base mount. There are so many different types of handle attachments that it's difficult to prescribe a standard of construction. Keep in mind the cross-section drawing of

FIGURE 4



Winding setup using an electric drill and Variac.

Figure 3. You'll need to mount a 3/8-inch diameter steel or brass bolt with no. 24 thread 1-1/2 to 2 inches long in the base after cutting off the head of the bolt. Use two-part epoxy to hold this bolt in place. The 5-minute kind will save construction time.

Now drill and tap through the bolt and handle to provide an electrical connection. Brass bolts (8-32) are ideal. Use a solder lug to attach the antenna wire, soldering to both the lug and the screw head. You should cover the complete base with shrink tubing or several coats of good enamel, or both.

Stem

Cut a length of wire, preferably no. 18 copper enameled (no. 20 or 16 will do), for the frequency desired. The length should be one-half wavelength plus approximately 12 inches. To find the length for one-half wavelength in feet, divide 468 by the desired frequency in MHz. You can do the winding by hand, with the help of another person, or by improvising a jig like that in **Figure 4**. If you have access to a 0 to 120-volt Variac, you can use a 3/8-inch or larger electric drill (see **Figure 4**). If necessary, use a handle with 3/8-inch or larger drill chuck mounted on it, instead of the electric drill setup shown.

With the shaft turning slowly, wind on about one-half or a little more of the wire allotted, starting with spacings of about 1 inch. As you reach the loading coil position, gradually reduce the spacing until the turns are barely separated. The bottom of the loading coil should be roughly 12 inches from the top of the pole.

Loading coil

The loading coil offers several variations. You are, of course, dealing with low power here — in the area of 100 watts or less. This is definitely not a high powered linear amplifier.

If you're using a "fat" pole (5/16 inch or larger diameter at the loading coil position), you can wind the loading coil directly on the pole — especially for a 10 or 12-meter antenna. However, VSWR usually seems to be better with a larger diameter. A loading coil of 1/2-inch PVC pipe works well, but if fiber glass poles with larger ends are available (say 1/2 inch or slightly larger) a short section cut from the larger end would be lighter and better.

You'll have to improvise when mounting the coil form. Corks with a hole in the center and some epoxy will do the trick. Cut or grind off the excess part. The small mushroom-shaped wooden buttons used for covering screw holes in furniture work nicely for 1/2-inch PVC by enlarging the hole in the end of the pipe. These buttons usually cost a dime or less at most hardware stores.

If the loading coil is approximately 1/2 inch in diameter, it should be about 3 inches long for 10 and 12 meters, 4 inches for 15 and 17 meters, and about 5 inches for 20 meters. Wind all the remaining wire, except for the last 12 or 14 inches, close spaced on the loading coil form. The remaining 12 inches or so of wire can be loosely spiraled up the pole to hold the wire steady for testing. You can use tape or epoxy to hold turns in place where needed, like the beginning and end of the loading coil.

Now you're ready to screw the assembly onto the spring magnet mount. Place the antenna on the ground plane in the yard or on top of the vehicle. Use a grid-dip meter and VSWR test equipment as described in the section on testing.

It may be necessary to cut off half or more of the last 12 inches of wire. If, after cutting it all off, the resonant frequency is too low, remove a few turns from the top of the loading coil and extend the last 12 inches as before. Then recheck for resonance with a grid-dip meter. This may all sound like a lot of guesswork, but it seems that if you've used the prescribed amount of wire, the antenna will almost always work out when you vary the last 12 inches. All the bands can have nearly the same pole length of about 5 feet, if you desire. The big variable is the loading coil. Remov-

ing 1 inch from the tip will make a rather large difference in resonant frequency, but you'd need to remove several inches from the loading coil to make a similar difference.

Tunable tip

You can make the tip adjustable if you wish by using telescoping sections from TV rabbit ears, old auto or portable radio antennas, and so forth. Use only two sections; choose those that will fit the rod above the loading coil after cutting off any excess fiber glass rod. Epoxy the tip onto the rod after sanding off a spot of chrome to allow for easy soldering of the wire end. The total midadjustment length of the telescoping section will be a little shorter than the length of wire it replaces because of the greater capacitance of the tubing. Recheck the VSWR, and if the midrange setting hits the desired VSWR at the frequency selected, the section is ready for completion — after you plug the hole in the end of the tip.

With the exception of the sliding tip, you can use heatshrink tubing over the length of the antenna. This can be a bit difficult because of the larger diameter of the loading coil, and possibly the base. It may be expedient to shrink tubing over the stem area before sliding on the loading coil form. Have two or three sizes of heatshrink tubing on hand. If places remain where the tubing doesn't fit closely, fill the gap with epoxy and sand smooth.

To shrink the tubing on long items like this antenna, try a toaster oven. This is another easy-to-find garage sale item; you can often find one for $3 to $5. You can dedicate your oven to this type of service by cutting a 3 or 4-inch diameter hole in the back. Finally, spray the antenna with two or three coats of good enamel.

If you're not particular about its bulky appearance, 1/2-inch PVC pipe makes quite satisfactory antennas. Using PVC eliminates the need for a form for loading coils. Otherwise, follow the procedure for coil winding and use the same amount of wire. The tunable tip will require some kind of cap or plug with a hole in it. Seal the tip well to keep water out.

Conclusion

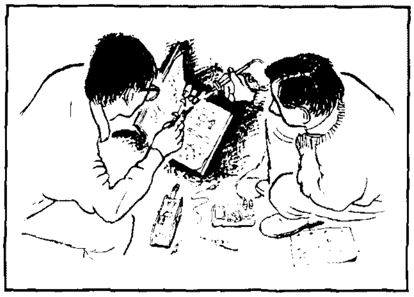
Anyone who likes to experiment will surely enjoy using some combination of the antennas described. On a recent trip to southern Missouri, I made contacts on all five bands using the antennas in Photo A and a Yaesu FT-101E modified for the WARC bands. I also worked 17 stations in Europe and Asia in 2 hours operating time. I made contacts on 40 meters by attaching a 33-foot wire to a common feedpoint and throwing the other end up into a tree. Attaching the 40-meter wire didn't affect the operation of the five helicals on the spring mount. Your results can be equal to those obtained with expensive commercial products.

The features I've tried to stress here are:

- Top loading has an advantage over middle or bottom loading.
- Top-of-vehicle mounting is superior to bumper mounting.
- Multiband operation is possible with one lead.
- *Good efficiency occurs at low power.*

But in the final analysis, great economy and the fun of building one's own equipment are probably the best features of all.

Design Curves For Loudness-compensated Volume Controls



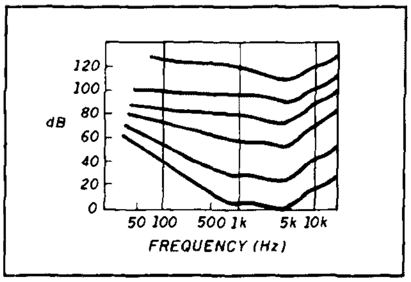
Hams deal with AF problems on occasion. Once one of my ham friends complained about the trouble he had adjusting the loudness-compensated volume control in his homemade FM receiver. I decided to analyze and test a popular loudness-compensated volume control circuit and make a set of experimental curves demonstrating how and to what extent different components' values affect its performance.

This work, though tedious, is worthwhile as one can easily determine the proper values of circuit components in minutes.

It's known that the human ear is not equally responsive across the audio range. It's more sensitive to mid-frequencies than to the lower and higher frequencies. This situation becomes more pronounced at low volume levels.

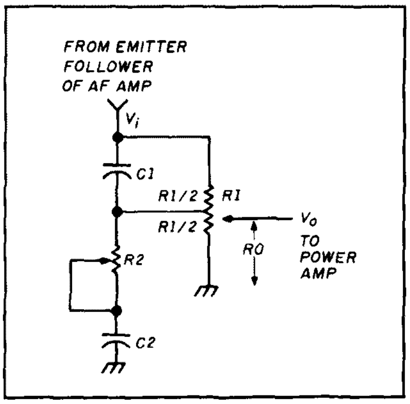
The equal loudness curves given in **Figure 1** show the nonlinear frequency response of the ear. We see that for equal loudness the amplifier output at bass and treble frequencies should be greater than at mid-frequencies. So, for best reception of music or other signals with a wide frequency range, it's desirable to use a volume control with properly designed loudness-compensated characteristics. This means at low volume settings the amplifier output should be greater as frequency goes lower or higher from the mid-frequencies.

FIGURE 1



Equal loudness curves to human ear at different power levels.

FIGURE 2



Loudness-compensated volume control circuit.

Circuit and working principle

The loudness-compensated volume control which I analyzed and tested is shown in **Figure 2**.

R_1 is a center-tapped pot. The combined impedance of C_1 in parallel with $R_1/2$, or R, decreases as frequency increases from the mid-frequencies, so the treble range is boosted. The combined impedance of C_2 in series with R_2 increases as frequency decreases from mid-frequecies, so bass frequencies are boosted.

Voltage transfer function of the loudness-compensated volume control

The voltage transfer function in dB of the loudness-compensated volume control through circuit analysis can be expressed as

A Digital Field Strength Meter

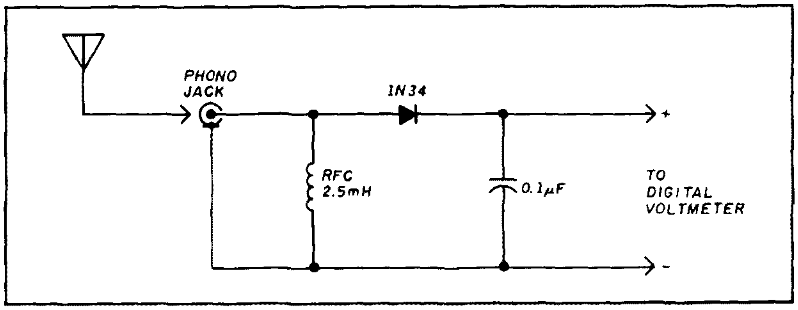
Here's an adapter that turns a digital voltmeter (DVM) into a field strength meter (see **Figure 1**). The DVM is well suited to the purpose because it's sensitive (down to 1 mV) and can detect slight changes in signal level (e.g., 0.1-percent change at 1 volt).

I built the adapter in a plastic film can. Banana plugs are epoxied into holes in the can so that the device plugs directly into the DVM. In operation the meter is set to the 2-volt DC range. The antenna is a short wire whip.

Parts aren't critical. The capacitor (0.1 μF) is larger than usually used. Unlike a conventional meter, the DVM has no inherent damping action; if its input has an AC component, it refuses to settle on a definite reading. The larger value capacitor reduces this problem.

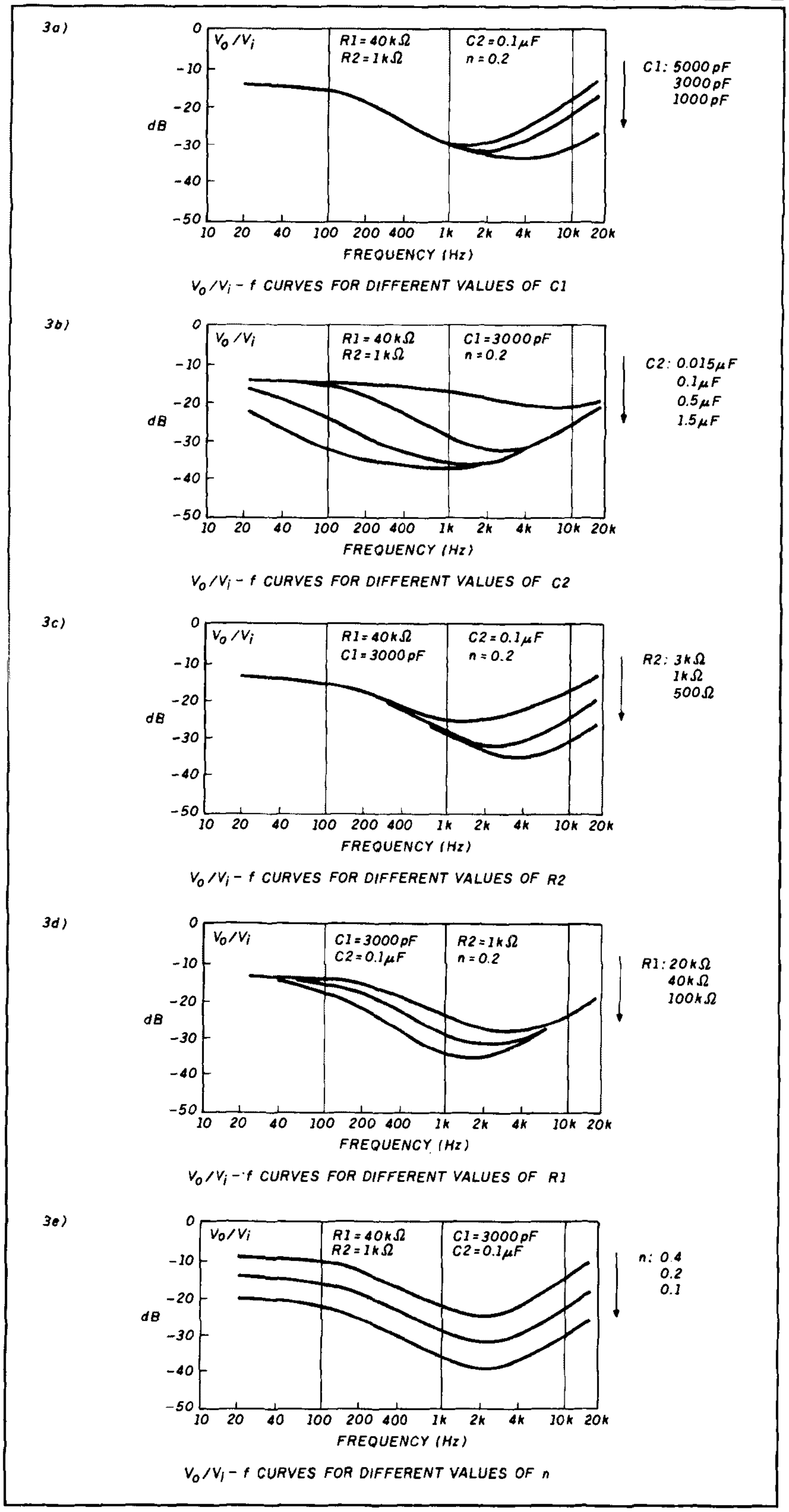
Michael A. Covington, N4TMI

FIGURE 1



Schematic for a digital field strength meter.

FIGURE 3



Voltage transfer function characteristics of the loudness-compensated volume control in *Figure 2*.

$$V_o/V_i = 20\log\left[2n\sqrt{\frac{A^2+B^2}{C^2+D^2}}\right] \qquad \textbf{(1)}$$

where

$$A = RR_2 - \frac{1}{4\pi^2 f^2 C_1 C_2}$$

$$B = R/2\pi f C_2 + R_2/2\pi f C_1$$

$$C = RR_2 - 1/2\pi^2 f^2 C_1 C_2$$

$$D = R(C_1+C_2)/2\pi f C_1 C_2 + R_2/\pi f C_1$$

$$n = R_0/R_1,\ R = R_1/2$$

since we are only interested in the low volume case, for n = 1/2.

Experimental curves and conclusions

The loudness-compensated volume control was tested for its voltage transfer function as a function of frequency under various circuit conditions. The experimental curves are shown in **Figures 3A** through **3E**.

All the experimental curves shown are in agreement with the computer data of the analytical expression, and give useful information about the effects of components values on the circuit performance.

These curves indicate that decreasing the value of C_1 will shift the minimum-output frequency (MOF) to the high frequency side and lower the output power in the high frequency range. The transfer function characteristics remain unaltered below mid-frequencies. Increasing the value of C_2 will mainly shift the MOF towards the low frequency side and lower the power output in the low frequency range. The transfer function characteristics remain basically the same above mid-frequencies if the value of C_2 is not too small.

Decreasing the value of R_2 shows somewhat the same effects as decreasing the value of C_1. Increasing the value of R_1 shows somewhat the same effects as increasing the value of C_2.

The following is a design example. With the component values shown in **Figure 3E**, the MOF is near 3 kHz. As the volume control is turned down, decreasing the value of n, both bass and treble are boosted.

Tseng C. Liao

SKYWAVE COMMUNICATIONS PART 3

Predicting skywave propagation

By Cornell Drentea, WB3JZO

Casual short term propagation predictions for HF communications can be made through simple observations of sunspot activity. Do-it-yourself types can make visual inspections of the sun's surface, using amateur telescopes equipped with dark glass to count the number of spots. The data can be input to the smoothed solar flux formula:

$$SF = 63.7 + 0.73\,R + 0.0009\,R^2$$

where R is the relative number of spots and can be obtained from:

$$R = k\,(10g + f)$$

where g is the number of sunspot groups,
f is total number of observed spots,
and k is the observer difference factor.

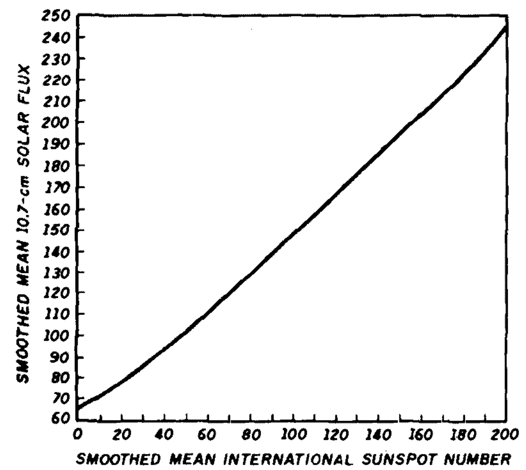
The smoothed sunspot number is then obtained by calculating the monthly mean relative sunspot number, Rm, which is the monthly average of all daily numbers. Additional smoothing is obtained by averaging all monthly numbers into a 12-month average called Rs (relative sunspot) expressed by the following formula:

$$R_S = \frac{\frac{1}{2}\,Rm1 + Rm2 + Rm3 + \ldots + Rm12 + \frac{1}{2}Rm13}{12}$$

When plotted, the smoothed solar flux value is an almost linear function of the smoothed number of sunspots as shown in **Figure 1**.

Sunspot observations started with Galileo in 1611. Recorded observations and the discovery of the 11-year cycle have been attributed to Hendrik Schwabe, a German pharmacist in the mid-1800s. Solar flux has been measured since 1947 by a Canadian observatory in Ottawa (readings are taken daily at 1700 UTC). Most of today's methods of measuring the sun flux use radio flux readings. In 1942 radar defense stations in Great Britain found that the relative RF noise amplitude coming from the sun was representative of the sunspot activity. It was later decided that the radio noise received at 2,800 MHz (10.7 cm) correlates almost linearly with the number of sunspots present on the sun (as was discussed earlier). This is the preferred method of observation today. Although it's possible to build simple amateur radiometer receivers at this frequency, the information can also be obtained from the WWV* broadcasts at 18 minutes after the hour as part of the geophysical alerts (geoalerts). WWV can be heard on 2.5, 5, 10, 15, and 20 MHz. Additional information in the bulletins deals with the estimated A-index, a measure of the earth's geomagnetic activity as impacted by the slowly arriving low energy protons

FIGURE 1



The solar flux is almost a linear function of the number of sunspots. In general, the higher the number of sunspots, the higher the probability of propagation through reflections and refractions in the F layer.

*WWV is operated by the Department of Commerce, National Oceanic and Atmospheric Administration, Environmental Research Laboratories in Boulder, Colorado. The stations are maintained at Ft. Collins, Colorado, and Kauai, Hawaii.

and electrons. This data is collected daily at a midaltitude site, like Fredericksburg, Virginia. The K-index gives additional data. This is the value of geomagnetic activity taken at 3-hour intervals, given for Boulder, Colorado. The K-index correlates closely with the A-index, as shown in **Table 1**. **Table 2** shows the relationship of all elements in a WWV geoalert. The short term probability of propagation has been graded from 0 to 5, with 5 being the best.

Here are some real life examples. A typical geoalert announcement may sound something like this:

"The solar flux is 235
The A-index is 10
The K-index is 2"

What does this mean? From the information in **Table 2**, you'll see that with a solar index of 235, skywave propagation should be very good at the F layer level. However, with an A-index of 10 and a K-index of 2, **Table 2** indicates that the geomagnetic field is *unsettled*. This makes for a grade 4 short term probability (good to fair) of propagation.

Another example may sound like this:

"The solar flux is 221
The A-index is 4
The K-index is 1"

Looking at **Table 2**, you can see that the solar flux is high enough to allow for good to excellent conditions. With an A-index of 4 and a K-index of 1, a period of quiet geomagnetic activity is expected along with grade 5 (excellent) conditions. I chose the preceding two examples as actual real life situations and have verified them with expected results on the air. **Table 2** should be a useful tool to give you a good feel for short term propagation. A very simple program could be devised for a programmable calculator or a computer using the algorithms shown.

In addition to these tools, the relative signal strength of WWV can also be a good indicator of immediate propagation conditions. A scan of these frequencies can reveal with some certainty what the MUF is at a given time.

TABLE 1

Correlation between the A-index and the K-index

A index	K index
0	0
3	1
7	2
15	3
27	4
48	5
80	6
140	7
240	8
400	9

Electromagnetic storm prediction

How can a geomagnetic storm be forecast? Immediately following a sun disturbance, a number of spots and consequent flares result. The number of flares varies from about 60 during the low part of the 11-year cycle to about 380 during the peak years. The flares produce a certain amount of ultraviolet and x-ray radiation which interacts with the ionosphere within 10 minutes of the outburst. This energy, together with the higher energy protons and alpha particles, plays an immediate role in ionizing all layers and providing

TABLE 2

Determining propagation from WWV. The propagation depends on the close relationship between the solar flux, the A-index, and the K-index, as shown. The probability of propagation is graded from 0 to 5 as follows:

0 = BLACKOUT
1 = VERY POOR
2 = POOR
3 = FAIR
4 = GOOD
5 = EXCELLENT

1. If solar flux $\leq$ 150, probability of skip is poor to good.
2. If solar flux $>$ 150, probability of skip is good to excellent.
3. If the solar flux index is as in notes 1 and 2, and the A and K-indices are as below, propagation probability is:

If A-index is:	And K-index is:	Geomagnetic activity is:	Propagation probability grade is:
$\leq$7	$\leq$2	Quiet	5 +
>7 but <15	$\leq$3	Unsettled	4
>15 but <30	$\leq$4	Active	3
>30 but <50	=4 or 5	Stormy (minor)	2
$\geq$50	>6	Stormy (major)	1
$\geq$100	>7	Stormy (severe)	0

good skywave communications — at least for a short time. If the solar storm is of sufficient intensity, the effect of the delayed particles discussed earlier produces quick recombinations in the D layer to a point of total RF energy absorption. This phenomenon usually happens 18 to 40 hours after *the initial flare outburst and results in* a total communications blackout. When this occurs, it is said that we are having a magnetic storm. The aurora phenomenon happens at the same time. In general, during periods of peak sun activity, a strong propagation outburst almost always culminates with a radio communication blackout.

Other tools for predicting skywave propagation

Among the more popular tools for predicting skywave propagation for the radio Amateur is a worldwide network of radio beacons on the 14 and 28-MHz bands. The most popular system operates at 14.1 MHz. Several beacons from various parts of the world transmit on the same frequency 24 hours a day at predetermined time intervals. They are time synchronized so that when one station ends its transmission, the next one starts up. To begin, each station identifies itself at a preset power of 100 watts for 9 seconds. Then a 9-second carrier transmission follows at 10 watts, another one at 1 watt, and another at 0.1 watt. The station resumes with a last 9-second transmission of identification at the original 100 watts. The next station in the cycle follows using the same procedure until all eight stations recycle.

By knowing the stations' QTHs and the particular power transmitted in the cycle, you can establish what direction the skywave propagation favors and the amount of power required to work in a particular direction. The stations are as follows:

Cycle no.	Call	Location
0	4U1UN/B	United Nations, New York
1	W6WX/B	Stanford University, California
2	KH6O/B	Honolulu, Hawaii
3	JA2IGY/B	Mt. Asama, Japan
4	4X6TU/B	Tel Aviv, Israel
5	OH2B	Helsinki, Finland
6	CT3B	Madeira Island
7	ZS6DN/B	Transvaal, South Africa

Each Station Cycle:

100 W ID	10 W CAR	1 W CAR	0.1 W CAR	100 W ID
9 Sec	9 Sec	9 Sec	9 Sec	9 Sec

While this method is an ideal indicator for propagation forecasting, it is unfortunate that HF packet has taken up the spectrum lately to well below 14.1 MHz (down to 14.095 MHz, to be exact), to the point that detecting these weak beacons may now require the most selective, crunch-proof receivers.

Computer programs

Various programs for predicting skywave communications are available for the radio Amateur and the professional alike. While most programs consider only the solar flux for determining propagation, a few use the A-index and several other parameters to determine the proper conditions. As I discussed earlier, long term prediction algorithms are generally used to overcome the day-to-day variations caused by the irregular behavior of the solar wind and its interaction with the magnetosphere and the ionosphere. Because solar disturbances are unpredictable in themselves, most long term programs concentrate on a median value of these variations.

The programs mentioned here come in *MS-DOS format* for the IBM and compatibles. Some programs are also available for other formats. Here's a brief examination of some of these programs.

MINIMUF is an effective simple program which uses the *solar activity, input as either the* number of sunspots or as the solar flux, as its main parameter for determining the MUF. After you type in the month and day, the solar flux as given by WWV, and choose the target location from an optional menu, the program calculates and plots the MUF for that particular area on a vertical scale against a UTC time table. You can obtain a printed copy of the information indicating all data; this also includes the coordinates of your location (if properly entered) and the coordinates for the target location, along with the bearing in degrees for the short path. The range to the target in statute miles is also given.

BANDAID is a menu-driven program with two versions — simple and fast — which takes advantage of an 8087 coprocessor. After a "HI" greeting in CW, the program asks you to make a choice between MUF, sunrise and sunset information, QSL information, beacon information, and several other options. A graphic feature can quickly pinpoint the QTH of an unknown call on a world map. Solar flux information is simple to input. The program uses the A-index in addition to the solar flux. The output is a horizontal graph containing four plots as a function of time (local or GMT, and target). From top to bottom, they are:

1. Highest possible frequency, HPF
2. Maximum usable frequency, MUF
3. Frequency of optimum transmission, FOT
4. Lowest usable frequency, LUF

The plot also prints the target information, the bearing in degrees, the distance to the target in miles, the flux, and the A-index, along with a qualification of the ionospheric *conditions at the time. BANDAID is a good overall propa*gation forecasting program.

MUFMAP *will* draw a customized world map showing the probability of communications with all areas of the globe at a particular time as a function of frequency and your QTH. Although it takes a little time to create a map (approximately 4 minutes on an XT), several maps for predetermined time intervals can be saved to a disk, and a MUFMOVIE of an entire period can be played back quickly. This program also *uses the solar flux to determine propagation.*

MINIPROP is one of the later programs. It is also menu driven, and accepts standard data like the solar flux. The date is directly input if the computer is equipped with time and date information. The program outputs an organized report showing coordinates, latitudes and longitudes of originating and target locations, and sunrise and sunset information. The report also includes *grayline information* for both locations, short and long paths in kilometers and miles, bearings from both QTHs, the number of possible

hops, and several other parameters. The next screen outputs expected signal levels in dB over 0.5 mV at the frequencies (bands) of interest. Other information includes UTC and MUF data. The probability of propagation is coded with four distinctive degrees of reliability for ease of use. They are:

A = 75 to 100 percent
B = 50 to 75 percent
C = 25 to 50 percent
D = 1 to 25 percent

From another menu you can chose to graph, zoom, or print the data. The graph shows MUF (on vertical axis) against UTC (on horizontal). A selection from yet another menu lets you model a particular band with signal level expectations in dB above the 0.5 mV level. The coded probability indicated above is used throughout. The program identifies about 360 prefixes and has provisions for modifying the entries. Another feature is a DX compass which shows, in graph form, MUFs for UTCs entered. In other words, you can look at this feature as showing direction of band openings at a particular time.

IONSOUND is a newly introduced software tool which uses the sunspot number (or 2,800-MHz solar flux number). In addition to predicting propagation, it provides total end-to-end link reliability calculations in varied situations. It does this by considering signal-to-noise ratios which include an atmospheric noise model at the receiver location derived from ESSA-ERL-110-ITS-8. Also calculated are radiation elevation angles, E and F layer as well as mixed E/F modes, propagation delays, bearing/path distances, and TX/RX antenna elevation gains. The minimum number of hops is calculated for a given takeoff angle, and signal strengths are calculated using free space loss. The program accounts for polarization losses in the ionosphere and receive antennas, in addition to ground reflection losses. E and F layer absorption losses are used to determine the critical vertical frequency for the F2 layer; this in turn is used to calculate the oblique F2 frequency. Total link reliability is obtained by multiplication of the uncorrelated path availability and signal-to-noise availability. The final result is a simulated chirp plot showing propagation nodes and their bandwidths, along with a tabular output summary. The summary shows prediction of propagation anywhere from 1.8 to 30 MHz by considering such parameters as transmitter power, receiver bandwidth, transmitter and receiver antenna gains, and local noise conditions. The software is relatively low in cost and includes a 26-page manual on disk. It comes in both coprocessor (8087/287/387) and noncoprocessor versions. Text is supported in various color option selections (CGA, EGA, and VGA graphics).

IONCAP stands for Ionospheric Communications Analysis and Prediction. The program was initially designed for a mainframe computer and is coded in FORTRAN. A PC version, IONPC2.5, is now available. This particular program is used for determining the reliability of given paths under various conditions, rather than for casual forecasting. IONCAP is different from other programs in that it is oriented totally towards probability. All parameters have statistical distributions associated with them. In addition, prediction errors associated with the distributions can be determined. This means you can cite any confidence interval for which you want the predictions calculated.

An interface software package called NECDEC provides compatibility with the popular antenna analysis program, MININEC 3 (MN). To some extent, the original program follows the ANSI 66 Standard. NECDEC is produced in modular form; this allows any subsection to be upgraded without affecting the rest of the work. It is divided into seven independent subroutines.

Input. The input subroutines handle the input options. Along with input information, they contain numerical coefficients for ionospheric parameters and atmospheric noise, as well as tables of parameters for skywave circuit performance. Optional antenna parameters can be input via the program, even if they are obtained from a source other than MININEC.

Path geometry. The path geometry subroutines determine the areas of the ionosphere to be sampled, and evaluate the magnetic field at the sample areas.

Antenna. The antenna subroutines evaluate antenna gains and output antenna patterns.

Ionspheric parameter. The ionospheric parameter subroutines evaluate (in detail) electron density profiles of the D, E, F1, and F2 layers. Absorption equations using an empirical modification of the secant law are used in this part of the program.

MUF. The MUF subroutine is based on an electron density profile derived from monthly median parameters of the ionosphere.

System performance. The system performance subroutines evaluate the critical circuit performance parameters with two basic subroutines — one for shorter distances and one for longer. Among other factors considered are the general noise at the receiving site as combined with signal-to-noise statistics which determine the reliability of a skywave circuit.

Output. The output subroutines generate all output options required by the program.

IONCAP provides complex and comprehensive predictions of communications systems which consider large directional antenna systems, diversity reception for slow data, FSK or facsimile, and high powered transmitters. The program is intended for those concerned with maintaining adequate service (signal-to-noise ratios) at given geographical points over sustained periods of time during any sunspot conditions. IONCAP is advanced enough to allow for applications like back-to-back reuse of the same frequency spectrum (broadcasting two independent transmissions on the same frequency, but beamed to two different places on earth without interference to each other) or other forms of diversity. Advanced Amateurs may find this program of interest.

Conclusions

This three-part series was designed to provide more insight into the phenomenon of skywave communications. I hope that you find the information presented timely. My purpose was to clarify some of the many misconceptions about the ionosphere and the way HF communications interact. I have made an effort to present information on useful tools for maintaining reliable communications for both Amateur and professional operations. Because of space limitations, certain areas like selective fading, propagation via meteor scatter, radiation angle, multiple hop, and tropo scatter have been intentionally omitted. These articles were intended primarily to fill a certain gap in the literature by presenting information not found elsewhere. I would like to

thank Dr. Mario Ierkic of the Arecibo Observatory, National Astronomy and Ionosphere Center; Dr. John Wolcott from the Los Alamos National Laboratory; Don Lucas, WØOMI, of Lucas Radio; and Jacob Handwerker, W1FM, for their inputs to these articles. Also many thanks to Bonnie Horishnyk; Rick Whiting, WØTN; and Earl Johnson, NØDTK, for reviewing this material. hr

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THE BI-SQUARE ARRAY

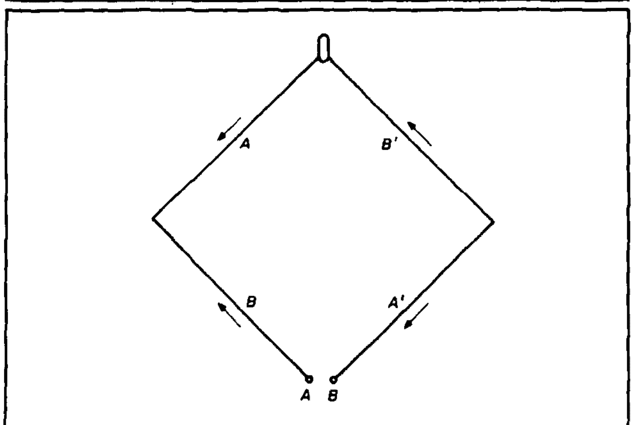
Old favorite works wonders on new 17-meter band

By C. Drayton Cooper, N4LBJ, P.O. Box 5, Bowling Green, South Carolina 29703

Interest in antenna experimentation has always been high among Amateurs. Now that the 17-meter band is open to United States operators, more and more of us are concerned with erecting antennas designed to operate on this band. As DX competition increases, we are becoming more concerned with building antennas that will provide some gain and directivity.

There are a number of us who don't have the resources needed to reproduce the great, world-beating antenna systems of ham legends like W6AM or W4BPD. These giants had acres of rhombics, V beams, and Sterba curtains. But there is an antenna that's functional on three bands, provides modest gain, directivity, is easy to put up, and requires only one support.

FIGURE 1



Because current flow reverses at half-wave points, elements A and A' are in phase. Elements B and B' are also in phase. In essence, this is a "harmonic antenna" which is end fed and cut to operate on the second harmonic. Essentially it is a 2-wavelength dipole folded back on itself.

The antenna is the bi-square array. Don't look for this antenna in the modern literature; your search will be in vain. Contemporary publications have long since dropped it from their repertoire, but the older books describe it in their sections on collinear, phased antennas.

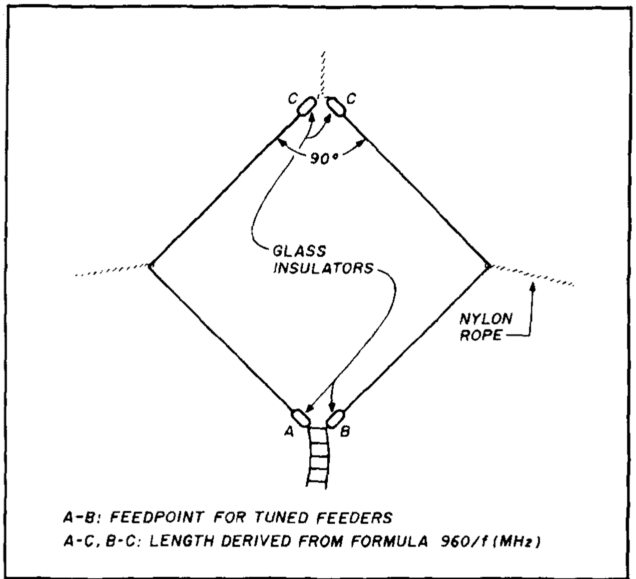
The bi-square antenna

Don't be misled by the commercial antenna with the same name. There is a company producing an antenna called the "bi-square," but it's a mini-bobtail curtain — not the traditional bi-square array. The traditional bi-square can be described briefly as a vertically polarized, 2-wavelength open loop, fed with open wire line or ladderline (see **Figure** 1). The antenna is somewhat more sophisticated than this description may lead you to believe. It operates as four collinear elements with each pair of elements in phase, yielding useful gain and broadside directivity on its design frequency.

It can be suspended from a central vertical support, much like an inverted V or "quad loop." The older literature presumed this to be a pole, or other nonconductive support. But most of us would hang it, as I have, from a tower. The fact that the tower is conductive seems to have little effect on my antenna.

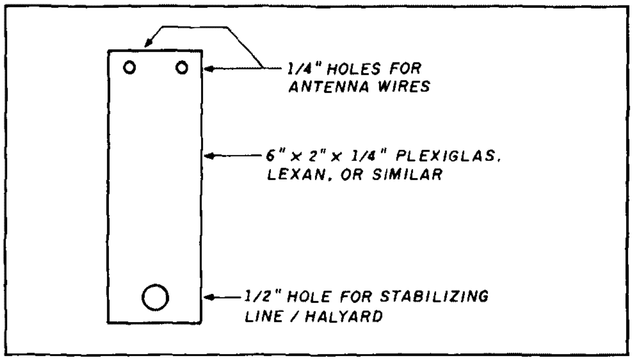
One of the more interesting features of the bi-square array is that it's useful over about a 2:1 frequency range. It's been my experience that my 17-meter bi-square performs very nicely on both 20 and 15 meters. I've made numerous DX contacts on all three bands. Signal reports from the southeast United States to Europe on 17 meters are running 1 to

FIGURE 2



Antenna performs best when feedpoint is about 1/4 to 1/2 wavelength above ground. Performance is adequate and acceptable at lower heights, down to about 3 to 5 feet above ground.

FIGURE 3



Good insulators can be made from a piece of Plexiglas™, Lexan™, or similar material cut to the dimensions indicated.

2 S-units better than they did on my 17-meter delta loop. I've noted a similar increase in signal strength on the African path as well.

I think it's safe to assume that the gain over a dipole would be on the order of 4 to 5 dB. Figures for the Lazy-H antenna, from which the bi-square array is directly derived, are on the order of 3.5 to 6.6 dB theoretical — depending on element spacing and height above ground. Bi-square array element spacing is on the order of 5/8 wavelength. If I'm correct in my assumptions, this would yield about 4 to 5 dB of gain.

Assembling a bi-square antenna

My self-supporting tower is 65 feet tall. At the top of the tower a yardarm extends out 3 feet on either side. A solid brass pulley with a halyard is mounted on this yardarm. In a vintage edition of *The ARRL Antenna Handbook*, I found the formula for element length for the bi-square antenna. Using this formula, 960/frequency (MHz), I cut two 53-foot lengths of antenna wire for a full wavelength at 18.1 MHz. I tied two glass insulators to the halyard and connected one end of each antenna wire to these insulators (see **Figure 2**). I secured nylon ropes to the halfway points of both wires to serve as spreaders, and to form the diamond configuration once the antenna is pulled into the air. At the bottoms of the wires, or feedpoint, I joined two more glass insulators with a short piece of nylon rope. **Figure 3** shows homemade insulators that you may use instead of glass insulators. Finally, I attached a stabilizing line to the bottom pair of insulators and pulled the whole assembly up. When I was satisfied with the overall "look" of the antenna — that it was squared up by the side ropes, and was hanging vertically and stable with the bottom stabilizing line fairly taut — I lowered it and attached the feedline.

This antenna requires balanced feedline. Years ago I used homebrew 600-ohm open wire, but today I use commercially available 450-ohm insulated ladderline. One side of the feedline is soldered to one side of the antenna; the other side is soldered to the opposite side of the antenna. When you've made these connections, hoist the antenna back into position. The bi-square array is ready to use!

Results

I was surprised to find that I received markedly better reports than I had with my previous antennas on the first day I used the bi-square antenna. I had been getting 559/569 reports from Europe on other antennas. Yet on the bi-square, broadside in that direction (as was my delta loop), the reports are consistently 569/599. Performance is good on 20 meters, but here the antenna becomes more of an "end fire" array with signal concentration in the plane of the antenna, rather than through it. I work Asiatic Russians (UA9, UAØ) easily with consistent 589/599 reports. The antenna also works well on 15 meters, but since that band has never been one of my favorites, I haven't run comparative tests on it.

As a passing note, I suggest you listen for N4SU on 30, 40, and 20 meters. Dave runs bi-square arrays on these bands with excellent signal strengths. His system consists of two arrays, 90 degrees apart, with a switching arrangement that lets him broadside his signals north/south or east/west.

It's also possible (if you have the room) to use bi-square arrays in a parasitic setup. A second array, spaced 20'6" apart (for 17 meters) from the driven array, is tuned either as a director or reflector. Several of the older antenna publications describe an L/C network to tune the parasitic array.

Once you've established parasitic element tuning for director/reflector status, you can insert a simple remote SPST switch in the network, controlled from the operating position, which enables the antenna to fire in either direction. A remotely switched tuning stub on the parasitic element could give you the same results.

The advantages of the bi-square array are its low cost, gain, directivity, and ease of assembly. And, should you use one, you can be certain of having a built-in topic of conversation on your contacts. Everyone you work, with the possible exception of some old-timers with long memories, is going to ask, "What in the world is a bi-square array?"

NONRESONANT DELTA AND V BEAM ANTENNAS

By Robert Wilson, KL7ISA, 115 Dogwood Trail, Leesburg, Florida 34748

Nonresonant antennas have great appeal. They're effective on all frequencies, simple, and inexpensive. The Army uses sloping V antennas for tactical operations because they're nearly foolproof. Vertical half-rhombic antennas are also part of the Army inventory, and are often advertised in ham journals. The Voice of America uses nonresonant rhombics and sloping Vs because of their wide flexibility. Amateurs can benefit from similar antennas, scaled down to fit a home owner's lot.

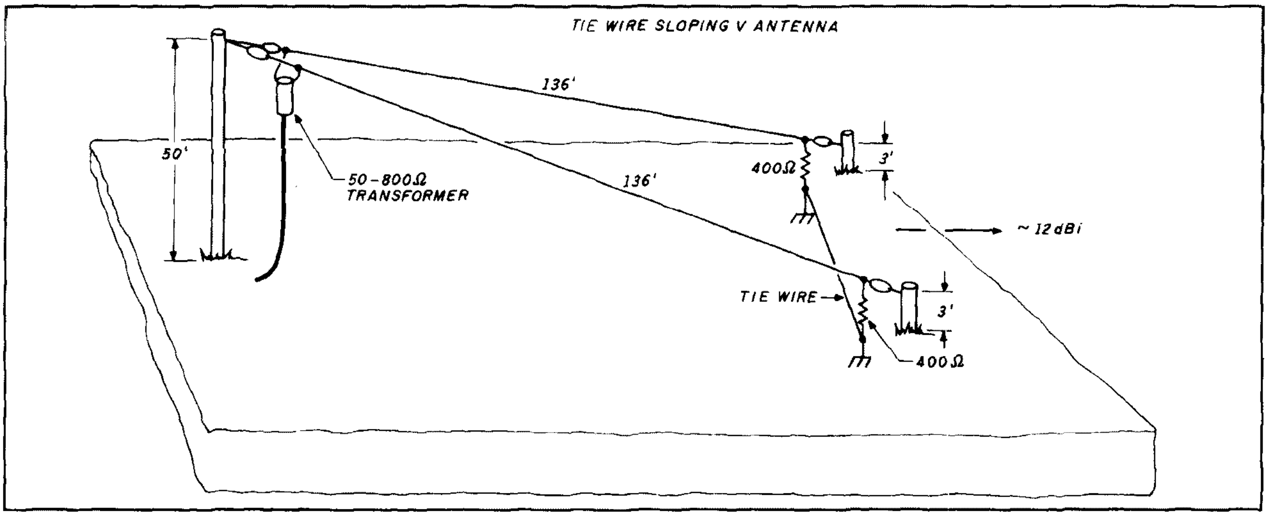
You can make nonresonant antennas that work on all bands without tuning from a few simple properly terminated long wires. These antennas are fed with 50 to 800-ohm RF transformers, and are often terminated with noninductive resistors. Sloping V antennas are commonly terminated with two 400-ohm resistors, each with a dissipation of one-fourth the transmitter power. Rhombics are terminated with a single 800-ohm resistor with a dissipation of one-half the transmitter power.

While working on commercial nonresonant V antennas, I noticed that in some parts of the world stations with these antennas had trouble with poor grounds. Other areas had a problem with people or animals running into the wires. I found two useful antenna variations that solve these difficulties.

Tie wire sloping V antennas

The tie wire sloping V is a basic sloping V antenna with an improvement. In general, V antennas slope downward from a single tall pole to two short end poles, and each leg is grounded through a 400-ohm resistor. A tested version of the sloping V antenna is shown in **Figure 1**. However, unless the soil is permanently wet, the grounds tend to give erratic results. The solution is both simple and elegant — simply tie the two grounds together with a wire buried a few inches underground. This tie wire stabilizes the ground effect for the dummy loads when the soil is dry. The wire

FIGURE 1



Tested version of the sloping V antenna.

TABLE 1

Comparison of some V and delta antennas optimized for 14.3 MHz, but useful for 1.8 to 29.8 MHz.

Type	Gain (dBi)	Leg (feet)	Stub (feet)	Width (feet)	Apex (degrees)
V	+9.4	58.5	DNA*	68.8	72
Delta	+9.3	58.5	34.4	68.8	72
V	+9.9	72.3	DNA	85.0	72
Delta	+9.9	72.3	42.5	85.0	72
V	+11.0	113.5	DNA	92.4	48
Delta	+12.4	113.5	46.2	92.4	48
V	+12.8	206.4	DNA	137.6	36
Delta	+12.0	206.4	63.8	137.6	36

*Does not apply.

TABLE 2

"Magic" leg lengths in half waves for delta and V antennas. (Other delta dimensions are not recommended.)

Leg length (half wavelength)	Delta Gain (dBi)	V Gain (dBi)	Comments
1.7	9.3	9.4	Compact antennas
2.1	9.9	9.9	Compact antennas
3.3	12.4	11.0	Best delta tradeoff
6.0	12.0	12.8	
7.5	12.8	13.6	
8.7	12.9	13.5	
10.0	13.1	14.9	Highest gain size

has almost no effect on the pattern because it's embedded (more or less) in the reflection plane of the antenna.

Delta V antennas

While looking for proof of the tie wire effect, I developed several other antenna variations. One of these was the delta V antenna. The delta V is an elevated V beam, with the two ends coupled together through a third radiator element or tie wire. This section of the antenna has a single 800-ohm termination resistor in the center. Such a V beam variation has several advantages over a rhombic. It requires only three poles, fits on small lots, produces nearly as much gain as a V, covers a wide frequency range, is independent of local ground conditions, is virtually foolproof — and people won't run into the wires! **Table 1** shows a comparison of two similar standard V and delta V antennas. Note that the delta V has a unique property — the gain peaks sharply for certain "magic" leg lengths where all of the patterns and phases work together. These peaks are shown in **Table 2**. The most desirable small sizes are 1.7, 2.1, and 3.3 half wavelengths per leg.

Calculations

The free space field of such an antenna can be produced quickly using a BASICA program.* This program optimizes the apex angle for one band, but it's possible to use the same antenna from 160 to 10 meters. The best results, and the most gain, are achieved with the antenna placed at about one half wave above ground. There may be an increase in indicated free space gain, shown by the computer program, due to reflection of the antenna image from the ground. This will be slightly different for every location and installation.

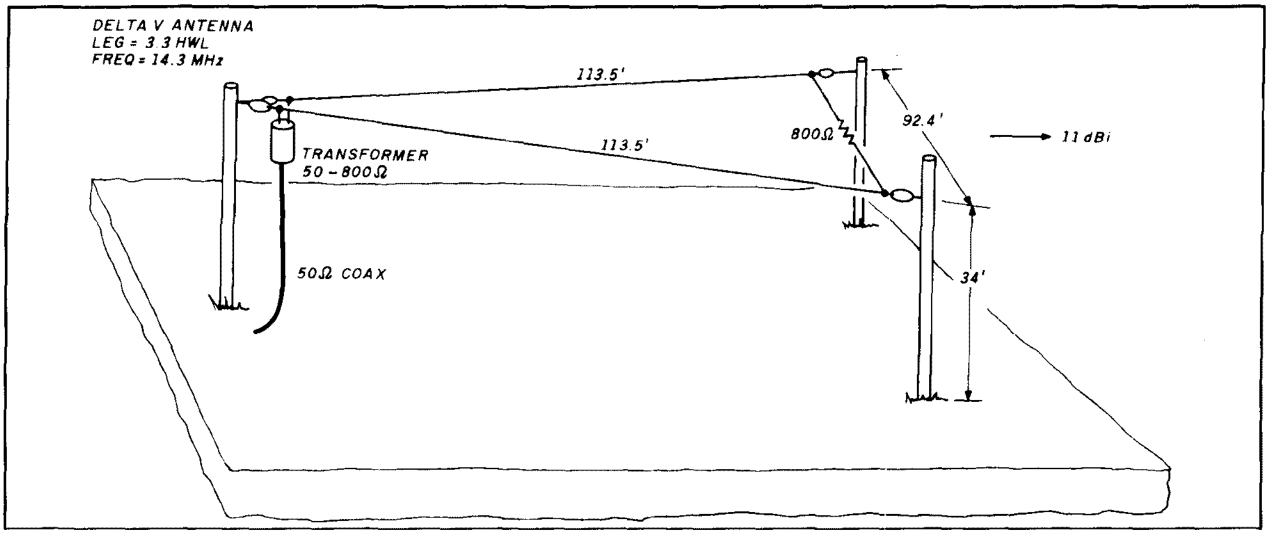
Program information

The program first calculates the two-dimensional free space pattern for a nonresonant wire of any selected length, given in half wavelengths. It then searches for the peak of the major lobe of this wire and gives you the value. You aren't required to use the free space optimum value calculated by the computer, but can select any value up to 89 degrees. Smaller angles may be useful for fitting the antenna to narrow lots, but there will generally be a loss of gain. Scaling the lengths to other bands and using the originally selected angle allows derivation of a pattern for any other frequency. The myth that rhombic-type antennas give high gain in one direction for all frequencies quickly evaporates when you're playing with V and rhombic programs. Instead, the patterns tend to split. However, these antennas will work effectively in one direction over at least a 2:1 frequency span, and will radiate reasonably in a forward direction over about a 15:1 frequency range. They tend to load well and often do not require tuners for the entire span. For radio Amateurs 14 MHz is a good center design frequency, while 11 MHz is excellent for commercial use.

After you set the desired angle the program rotates, calculates, and stores the vector sum of the pattern from the two wires. This is a theoretical free space V beam pattern with a phase center at some point between the two antenna wires. For the delta V, the program calculates the radiation pattern for the basic end stub wire. It rotates the stub pat-

* Program listings 1 and 2 are available from *Ham Radio* Magazine for an SASE with 25-cents postage.

FIGURE 2



Delta V antenna optimized for 20-meter band. Antenna requires pie shaped lot size of 104 feet by 93 feet.

tern of one wire, adds it to the pattern of its mate, and stores the pattern that results. This pattern also has a common phase center.

The feed phase delay between the two antenna systems is calculated for the delta V and the two combined patterns are added by vector methods. The resultant pattern is normalized to an isotropic value and decibel gains of the field are calculated for each angle. As a result, the field radiated by the end stubs of a delta V antenna is largely overcome by the V element field, particularly if you select certain leg lengths.

Dimensions

Figure 2 shows a possible city lot-sized delta V antenna optimized for 20 meters, requiring two V legs 113.5 feet long. Each stub leg is 46.2 feet, for a total spacing between the front supports of 92.4 feet. An antenna of this size needs a V-shaped space 104 feet long by 93 feet wide. Antenna height should be about 34 feet. The apex angle of the V is 48 degrees. The two stubs are joined in the center by a single noninductive 800-ohm resistor. Loading should be adequate from 160 to 10 meters, depending largely on the quality of the transformer or 800-ohm feed system.

Dummy load

You can make the 800-ohm load resistor by paralleling fifty 39-k, 2-watt carbon resistors. Solder the resistors between two parallel copper wires for an 800-ohm, 100-watt noninductive resistor. If you use this resistor only on single sideband, slightly more than 200 watts may be sent to the antenna because half the power is radiated away before it ever gets to the resistor. I found that average output powers of more than 200 watts for more than a few seconds will permanently change the value of the resistors.

RF transformer

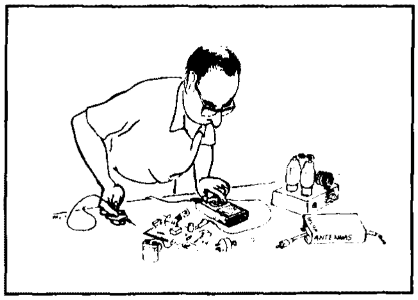
Make the transformer from a quality high frequency powdered iron RF toroid core. Not all cores are good for high frequency transformers. Use only RF quality wideband power cores with cross sections of at least 1 by 1 cm and a diameter of about 5 cm. Palomar* part F-240 Mix 61 or T-200A Mix 2 are adequate up to 400 watts. Core T-400A Mix 2 should be good up to 1 kW. Wind this toroid core with 20 turns of nonoverlapping insulated wire to make an 800-ohm secondary. The best wire is no. 18 AWG Teflon™ covered stranded wire, but it's costly. You can wind the five-turn 50-ohm primary over the top of the secondary using the same or larger diameter wire. Distribute the primary evenly over the entire core. Twist the 50-ohm wire pair together using about two to four turns per inch to identify them and make sure they won't radiate. Leave the 800-ohm pair well separated to avoid arcing at high power and undesirable high impedance effects. It helps to have the two sets of wires come out on opposite sides of the toroid. For a more permanent (but not so pretty) transformer, secure the ends in place with nylon tie wraps and cover the entire transformer with a heavy coat of clear silicon glue. Expect to wait a week for all of the silicon to harden completely.

Conclusions

The delta V antenna should work well in one direction for the selected band, and should be entirely effective from 160 through 10 meters and other services in between. The regular tie wire V will tend to give more uniform gain over a wider band, but you may need to put fencing at the lower wire ends to keep people away. Remember that the apex (or pointed end) of the V points away from the target, not toward it. You'll find it necessary to use a computer great circle azimuth calculation program or great circle map to point your antenna correctly. If you use a compass, and correct properly for local magnetic declination, you should be right on target. I'm sure you'll like the simplicity, uniformity of operation, and all-band capability of the delta V and tie wire sloping V antennas.

* Palomar Engineers, PO Box 455, Escondido, California 92025. Phone: (619)747-3343.

Ham Radio Techniques



MAY PERAMBULATION

By Bill Orr, W6SAI

Ten meters has been great! Though the band will drop out during the coming summer months, it will roar back in the fall. As we "roll over" the top of the current sunspot cycle, and observe a gradual decline in sunspot numbers, past experience tells me that we'll experience fewer solar blackouts and fewer disturbed days like those melancholy periods that obliterated 10 meters during the last year. Ten meters should be a red hot band for the next few years, and that's nice.

What exciting opportunities the band offers! In addition to the traditional CW, SSB, RTTY, and SSTV, 10 meters offers packet radio, satellite operation, NBFM, and also AM. AM??? Yes! There's quite a bit of AM operation between 29.0 and 29.2 MHz.

I must admit that a good AM station, with a well-designed class B modulator and a Western Electric 387W double button carbon microphone, sounds pretty good these days compared with the run of the mill SSB rig with a cheap dynamic microphone! Ah, nostalgia...

Yes, AM is alive and healthy. Listen around 29 MHz. If you want more information on this branch of our hobby, write to *Electric Radio* magazine, Box 139, Durango, Colorado 81302. It's a first-rate publication devoted to AM and old-time ham gear that can be operated on today's bands.

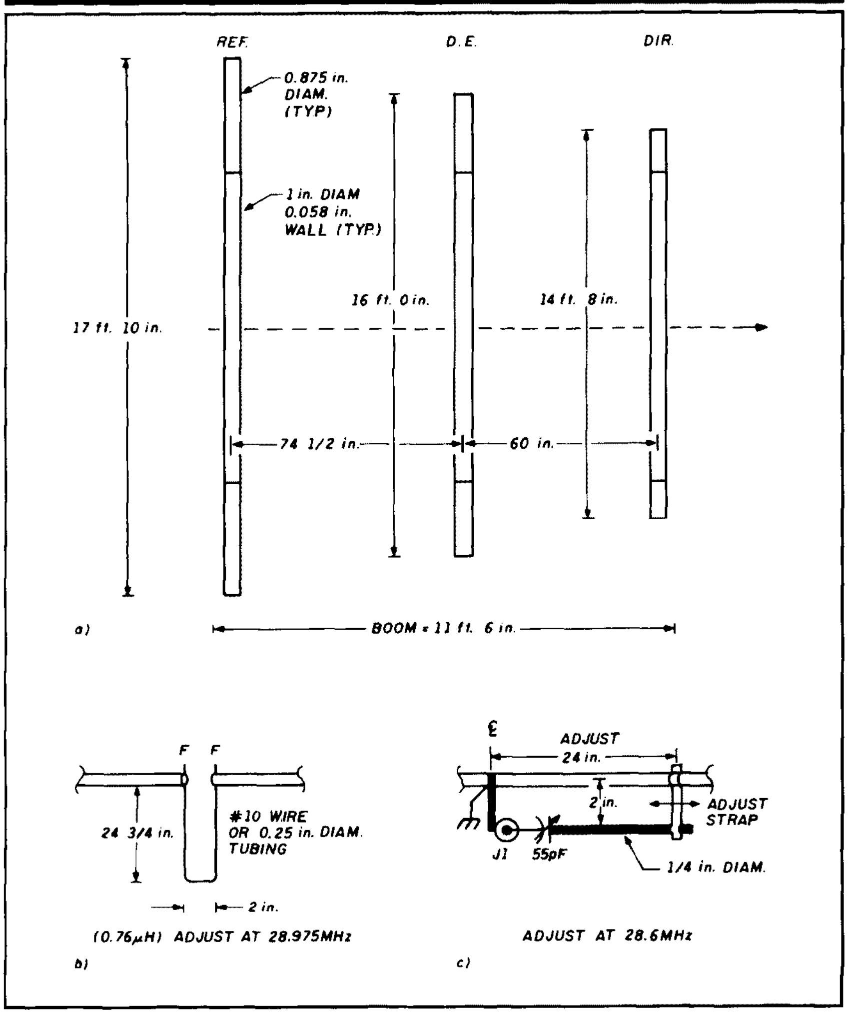
A broadband 10-meter Yagi beam

How about a three-element Yagi that will cover the whole 1.7-MHz wide band? That's a rarity. A Yagi tuned up at the low end of the band (say, 28.5 MHz) won't work very well at the top end if you want to try FM operation. The gain disappears, the front-to-back ratio deteriorates, and the SWR rises ominously. Sometimes the beam works better off the back than it does off the front!

The Yagi Optimizer (YO) program discussed last month* can provide an answer to this perplexing problem. It clearly shows that Yagi design is a mixture of tradeoffs. The secret of success lies in searching for the best combination of gain, front-to-back ratio, and

*The Yagi Optimizer program by Brian Beezley, K6STI, 507-1/2 Taylor Street, Vista, California 92084.

FIGURE 1



Beam is built on 2 × 0.065-inch tubing. Element center sections are 12 feet long (a). Hairpin match (with balun at F-F) provides SWR better than 1.75-to-1 across band (b). Gamma match doesn't require split driven element (c).

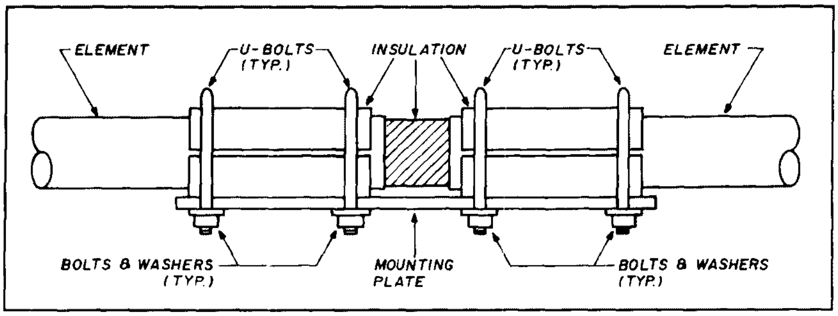
SWR that can be achieved within a given bandwidth. I've conducted such a search with the YO program and have come up with a practical Yagi that covers the whole 10-meter band. The essential characteristics are shown in **Figure 1**.

Forward gain is about 1.5 dB less than a narrowband design, but still a respectable 4.74 dBd at the design frequency. The front-to-back ratio averages 17 dB and the SWR across the band is below 1.75:1. Those figures are hard to beat!

Optimum SWR response across the band is achieved with a hairpin match and a split driven element. If a gamma match is used, SWR at the band edges rises to about 2.5:1. For those builders whose equipment can work into a SWR of that value, I've included dimensions for the gamma match section in **Figure 1**. The design is for 1.0-inch diameter antenna elements with tip sections of 0.875-inch diameter.

Elements are mounted atop a 2-inch diameter boom with U-bolts and a rectangular mounting plate 2 × 6 inches. The driven element is insulated from the bolts and plate by an insulating segment cut from PVC plastic water pipe (**Figure 2**). The two halves of the element are joined by a wood dowel plug sprayed with clear acrylic to protect it from the weather.

FIGURE 2



Driven element is split at center and joined by wood or phenolic dowel. Element is insulated from mounting plate and bolts by half sections of PVC water pipe cut to size.

Adjusting the match

The beam is designed for a free space environment. Ground effects may require that the match system be adjusted. Resonant frequency of the antenna is 28.5 MHz, but for lowest SWR across the band the match is adjusted at 28.975 MHz.

In the case of the hairpin match, the variables are the inductance of the hairpin and the reactance of the driven element. The driven element is cut about 6 inches shorter than resonance; the final length is given in the program's table. The hairpin length isn't very critical; usually the beam can be tuned up by merely "touching up" the driven element length.

The gamma match can be used but rod length, capacitance of the series capacitor, and driven element length are interlocked. The driven element is set to the dimension given in the program's table and the gamma capacitor adjusted for lowest SWR at 28.6 MHz. Rod length may have to be adjusted to bring the SWR down across the band. If a reasonable SWR response isn't found, the driven element is shortened an inch on each end and the test repeated.

You can get good results by adjusting the match when the beam is only 15 to 20 feet in the air. The perfectionist will want to "touch things up" when the beam is in final position atop the tower.

Propagation prediction

Propagation prediction was easy when I was a newly licensed Amateur. I would come home from school and look up at the evening sky to see if there was a full moon. I would note the low clouds on the horizon and sniff the air. I could determine quickly if there would be DX that evening. My pal, W2LX, improved upon my techniques by watching the sea gulls that floated over Long Island Sound. Between the two of us, we knew when 20 meters was open for DX.

Well, propagation prediction is much more complicated today. World War II demands for reliable intercontinental communication led to the development of a whole new world of propagation prediction that included such unlikely things as spots on the sun and the solar cycle.

That took a lot of fun out of DX prediction. But it provided more reliable results than sea gulls or the smell of burning leaves in the fall.

A new science grew up which concerned itself with solar predictions of one kind or another. In 1948 the National Bureau of Standards provided a cookbook procedure for propagation prediction, and programs in BASIC, FORTRAN, and RPN were gradually developed which handled a lot of the tedious mathematics of prediction.

In 1982 the MINIMUF program was developed by the Naval Ocean Systems Center at Point Loma, California. This made practical the prediction of the maximum usable frequency (MUF) for a given path on the home computer.

Sheldon Shallon, W6EL, followed the propagation programs closely and released MINIPROP 1.0 in 1985. This was the most comprehensive program for microcomputers available, and it was widely used by Amateurs and commercial radio station personnel. Version 2.0 of MINIPROP (1987) contained superior prediction techniques used by the British Broadcasting Company and other features obtained from information provided by the International Radio Consultative Committee of the ITU (CCIR).

The present version of MINIPROP, 3.03, provides better MUF and signal level prediction than the earlier versions, and has a lot of "bells and whistles" of great interest to ardent DXers and ragchewers.

MINIPROP 3.03

MINIPROP 3.03 uses a unique new method of finding the strongest ionospheric mode at times and frequencies of your choice between 3 and 30 MHz. It takes into account effects of the D, E, and F layers of the ionosphere. You enter the locations of two stations, the date, and the sunspot number (or solar flux), and MINIPROP does the rest.

What do you get? Predicted short and long path signal levels on frequencies of your choice, MUF, and E layer cutoff at two-hour intervals throughout the day. Band openings are noted and beam heading, path length, and sunrise/sunset times for the path terminals are given.

TABLE 1

Partial printout of MINIPROP great circle bearings from my QTH to 350 worldwide locations

Short Path Great Circle Bearings
From W6SAI 37.30 N 122.30 W
Add or subtract 180 degrees for long path bearings

Prefix	Location	Degree	Prefix	Location	Degree
FK	New Caledonia	242	JD1/MT	Minami Torishima	286
FM	Martinique	95	JD1/O	Ogasawara	294
FO	French Polynesia	207	JT	Mongolia	330
FO/A	Austral Islands	208	JW	Svalbard	8
FO/M	Marquesas Islands	203	JX	Jan Mayen	20
FO0X	Clipperton	153	JY	Jordan	19
FP	St. Pierre/Miquelon	57	KC4	Antarctica	180
FR	Reunion	7	KC6/E	Micronesia	271
FR/G	Glorioso Island	23	KC6/W	Belau	283
FR/J	Juan de Nova	45	KG4	Guantanamo Bay	100
FR/T	Tromelin	9	KH0	Mariana Islands	283
FS	Saint Martin	93	KH1	Baker, Howland Is.	247
FT8W	Crozet	157	KH2	Guam	283
FT8X	Kerguelen Islands	211	KH3	Johnston Island	257
FT8Z	Amsterdam/St. Paul	261	KH4	Midway Island	276
FW	Wallis/Futuna	236	KH5	Palmyra, Jarvis	235
FY	French Guiana	99	KH5K	Kingman Reef	240
G	England	33	KH6	Hawaiian Islands	251
GD	Isle of Man	33	KH7	Kure Island	276
GI	Northern Ireland	33	KH8	American Samoa	231
GJ	Jersey	35	KH9	Wake Island	275
GM	Scotland	31	KL7	Alaska	335

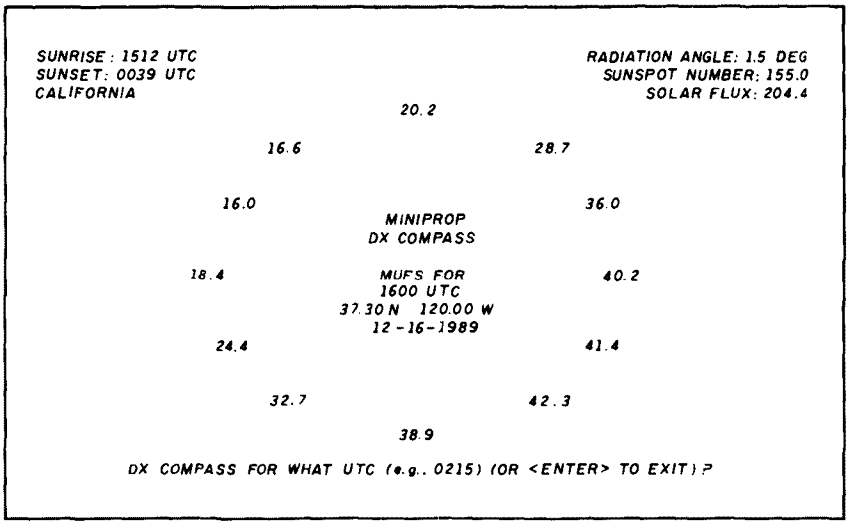
In addition, you get an atlas giving great circle bearings from your location to all atlas locations, grayline information and optimum antenna radiation angle for the paths in question. Finally, the program provides predicted signal levels for the five primary HF bands. In summary, it will do everything except work the DX for you!

To get started with MINIPROP you need to know the latitude and longitude of your station. If you don't, you can use the MINIPROP atlas of more than 350 locations. For California, the general latitude figure for computation is 37.3 degrees; the longitude is 120 degrees. For the precise location of your QTH, consult a geographic atlas or aeronautical chart.

I determined my location from an aeronautical chart. In seconds MINIPROP printed the great circle path bearings from my QTH to 350 worldwide locations (**Table 1**). Next, I entered the smoothed sunspot number (SSN) and printed a MINIPROP DX compass for my QTH (**Figure 3**).

I then asked MINIPROP to compute the path between my location and India. I quickly received the printout shown in **Table 2** giving the pertinent information. The short path prediction took the form of a graph (**Figure 4**). For

FIGURE 3



The MINIPROP DX compass for Northern California, December 16, 1989 shows MUF for twelve compass directions at 1600 UTC. Sunspot number of 155 was used.

0200 UTC, it seemed that 21 MHz was a good choice to work the elusive VU stations. A look at the signal levels for the period 0000 to 0530 UTC provided additional information.

The 15-meter band looked best around 0130 to 0200 UTC, and it appeared there might be a chance of 10 meters opening about the same time. But look at the 40-meter band! There's a long opening here, with reasonable signals from 0000 to about 0400 UTC.

You'll find enough good information in MINIPROP to amaze you. Armed with up-to-date solar flux data from

TABLE 2

"Quick Look" prediction for December 16, 1989 over California-India path. At 0200 UTC, 21 MHz looks like a good bet.

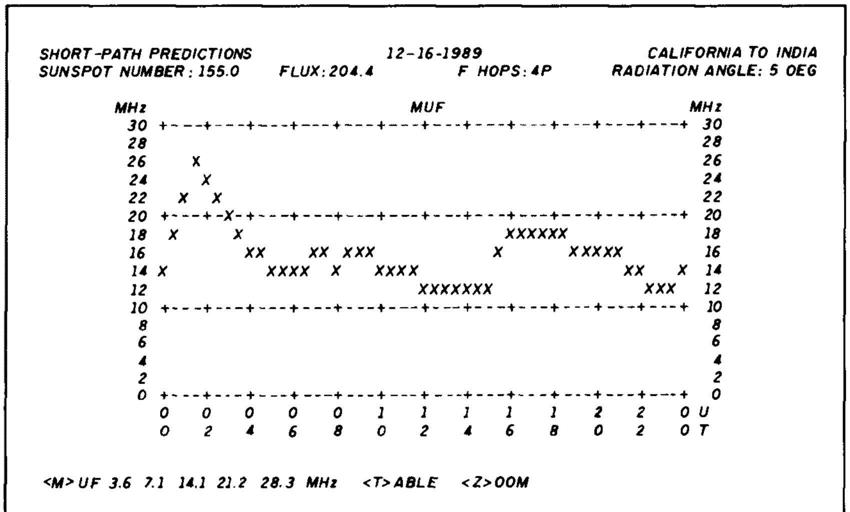
MINIPROP Short Path Predictions 12-16-89 Path Length: 12716 km
Sunspot number: 155.0 Flux: 204.4 F Hops: 4P Radiation Angle: 5 degrees
TERMINAL A: 37.30 N 120.00 W California Bearing to B: 340.2 degree
TERMINAL B: 25.50 N 80.00 E India Bearing to A: 17.4 degree
Terminal A Sunrise/Set: 1512/0039 UTC Terminal B Sunrise/Set: 0123/1148 UTC

Quick Look Signal Levels in dB above 0.5 mV

UTC	MUF	3.6 MHz	7.1 MHz	14.1 MHz	21.2 MHz	28.3 MHz
0000	14.5	23 a	28 a	26 B		
0030	18.0	28 a	30 a	27 A	24 D	
0100	21.7	27 a	29 a	27 A	24 B	
0130	26.3	24 a	28 a	26 A	23 A	21 D
0200	23.3	16 a	25 a	25 A	23 A	20 D
0230	21.3	8 a	22 a	24 A	22 B	
0300	19.8	1 a	19 a	23 a	22 C	
0330	18.5	−5 a	16 a	23 a	22 D	
0400	16.7	−10 a	14 a	22 a	21 D	
0430	15.4		13 a	21 b		
0500	13.7		11 a	21 c		
0530	14.0		11 a	21 c		

Signal levels not shown if below −10 dB or if predicted availability is zero. Availabilities A: 75-100 percent, B: 50-75 percent, C: 25-50 percent, D: 1-25 percent. a, b, c, d: Same as A-D, with high probability of reduced signal levels.

FIGURE 4



MINIPROP prints graph of California to India path for sunspot number of 155. X-axis is time in UTC. Y-axis is frequency in MHz.

WWV, MINIPROP provides you with the most accurate propagation information available.

Last fall I ran a test program from California to India using MINIPROP. Thanks to my MINIPROP data, I nailed Diego Garcia before the DX multitude descended! That made me feel very, very good!

MINIPROP will run on an IBM clone, PS/2, or true compatible with a 320K (or bigger) memory, one 5-1/4 inch floppy or 3-1/2 inch drive, and PC-DOS or MS-DOS 2.11 or later. An 80-column monitor (monochrome or color) is required. A math coprocessor is recommended but not required.

For more information, contact W6EL Software, 11058 Queensland Street, Los Angeles, California 90034-3029.

The Dead Band quiz

My March column posed two brain teasers. The first asked you to identify the quotation, "Remain on patrol in vicinity of Rockall."

What was this signal? To whom was it sent, under what conditions? What was Rockall?

The quotation is from *The Cruel Sea*, the outstanding book by Nicholas Monsarrat. The coded wireless message from the British Admiralty to the frigate "Saltash" signified the end of World War II. Rockall is a sea mount, about 300 miles west of Scotland. *The Cruel Sea* is recommended reading for a dead band!

The second quotation: "Brain! Brain! What is Brain?" is well-known to avid "Trekkies." It's from the third season of *Star Trek* in the episode "Spock's Brain." It was said by "The Leader" as the crew of the Enterprise struggled to release Spock's brain from "The Controller."

Finally, kudos to the following who also solved the "snowplow" problem: W5UOJ, KR4N, OH1AXV, PA3FCV, and W9ZSJ who found the problem in the book *Differential Equations* by Ralph P Agnew (McGraw-Hill, 1960). He says, "If you ever want to read differential equations for pleasure, try the book by Agnew." hr

INEXPENSIVE TRIBAND MOBILE ANTENNA

By Phil Salas, AD5X, 1517 Creekside Drive, Richardson, Texas 75081

It's not unusual today for many Amateurs to have VHF, UHF, and HF rigs in their cars. I have a Yaesu FT-727R dual band (144/450 MHz) handheld and a Cobra 146GTL CB SSB rig converted to 10 meters.

I've been using a shortened CB magnetic mount antenna for 10 meters. I mounted a cup holder on my dashboard (25 cents at Western Auto) and use this as a mount for my Yaesu 144/450-MHz rig. The small antenna on the handheld works okay through the windshield when I'm traveling around town. However, I've wanted an external antenna for the Yaesu so I can get more range. The only thing that stopped me from buying one of those dual band 144/450-MHz antennas was the thought of a third antenna (including my broadcast antenna) on my small car.

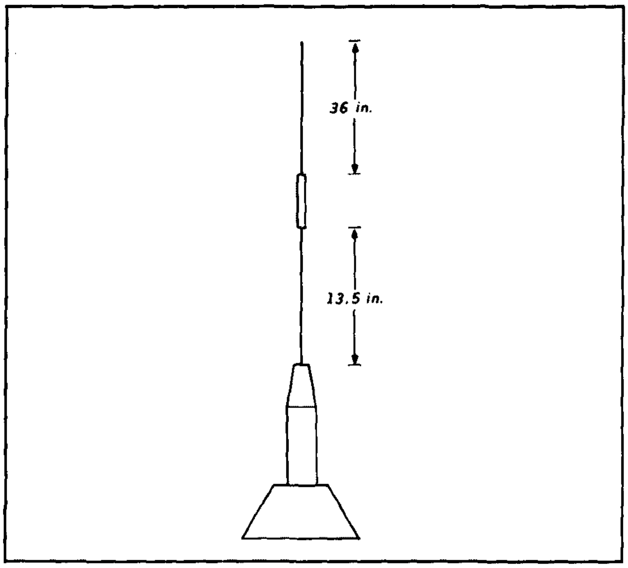
Radio Shack to the rescue! While wandering through my local Radio Shack, I saw a VHF/UHF/AIR magnetic mount mobile antenna. The Radio Shack part number is 20-018. For $26.95 I thought I'd gamble and see if I could make it work on my three mobile bands. Because the antenna cable has a Motorola plug on it, I also purchased the Radio Shack 278-117 BNC-to-Motorola adapter.

Tests

I took the Radio Shack antenna to work and swept it with a return loss bridge. The antenna showed resonances at 38, 150, and 450 MHz. This definitely showed some promise.

I next took the unit and attached it to the top of my car. By using a VSWR meter, I found I could achieve a 1.6:1 VSWR on 146 MHz and a 1.3:1 VSWR at 446 MHz when I adjusted the spacing between the bottom mount and the top resonator to 13-1/2 inches (see **Figure** 1). Even better, the measured VSWR of the VHF/UHF flex antenna supplied with the Yaesu was 2:1 at 146 MHz and 3:1 at 446 MHz.

FIGURE 1



New spacing between bottom mount and resonator provides VSWR improvement.

A labor of love

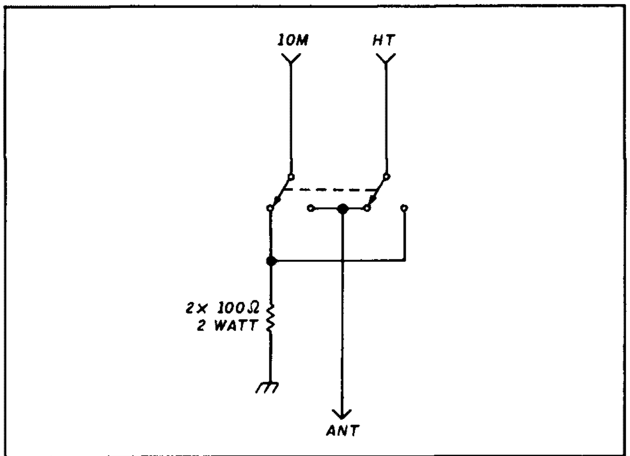
A quick calculation indicated that a 36-inch top section in place of the 14-1/2 inch section that came with the antenna should give me resonance in the 10-meter band. I purchased a 0.1-inch diameter, 36-inch long welding rod from my local hardware store and was able to achieve nearly a 1:1 VSWR in the 10-meter band. But hold on — it gets even better.

Because the trap apparently isolated the top section from the bottom at 146 and 446 MHz, I surmised that the bottom section must look like a quarter wave at 146 MHz and a three-quarter wave at 446 MHz. I removed the trap and top section, and found that I could still achieve the low VSWR on both 146 and 446 MHz when the bottom rod extended 14 inches above the bottom mount. To do this, I had to purchase a new bottom section (0.2-inch diameter aluminum rod), as the original rod was just a little too short. Again, my local hardware store had what I needed. Now when I attach the top section, the new bottom rod extends 1/2 inch into the trap, giving the desired 13-1/2 inch section between the bottom mount and the trap.

Switching arrangement

Finally, I replaced the set screw on the bottom end of the resonator with a 10-24 thumb-tightened screw. This permits me to remove the top section of the antenna easily whenever I don't need the 10-meter capability. An inexpensive antenna switch mounted on my dash now selects either the handheld or the 10-meter rig. I made my switch out of a Radio Shack 275-403 DPDT toggle switch mounted in a Radio Shack 270-235 2-3/4 × 2-1/8 × 1-5/8 inch aluminum box, two BNC female connectors, and an SO-239 coax connector. To keep from damaging the off-line rig should I accidentally key it, I also added a 50-ohm termination on the switch (two 100-ohm, 2-watt resistors in parallel) as shown in **Figure 2**.

FIGURE 2



Resistors in switch line prevent accidental damage if unused rig is keyed inadvertently.

Results

How does it all work? Great! I have terrific coverage on VHF and UHF, and I still work all over the United States, Canada, and South America (and occasionally Europe) on 10 meters with ease. Even with the longer top section, the magnetic base holds the antenna on my car roof at higher speeds. Not a bad deal for less than $35 worth of parts!

There you have it. This may be the simplest and quickest project around. You can find everything easily and inexpensively at your local Radio Shack and hardware stores.

Tom McMullen, W1SL

ELEMENTARY ELECTRONICS: ALTERNATING CURRENT AND CAPACITORS

In my April column I showed some of the interactions between inductance and alternating current (AC). There's a similar relationship between capacitors and AC as well. Let's look at how this works.

You've seen that when AC is applied to a resistor, the resulting current flow is right in step with the voltage. When the voltage is at its maximum value, the current is also at maximum, and so on. However, when an inductor is involved, the current gets behind the voltage. This is called current lag. Capacitors also separate the current and voltage peaks, but in the opposite direction; the voltage lags the current.

How does the current get ahead?

The current doesn't actually *gain* time, although at first glance it seems to do so. It's a relative thing, so when the voltage lags the current, it appears that the current is somehow getting ahead. Here's what happens.

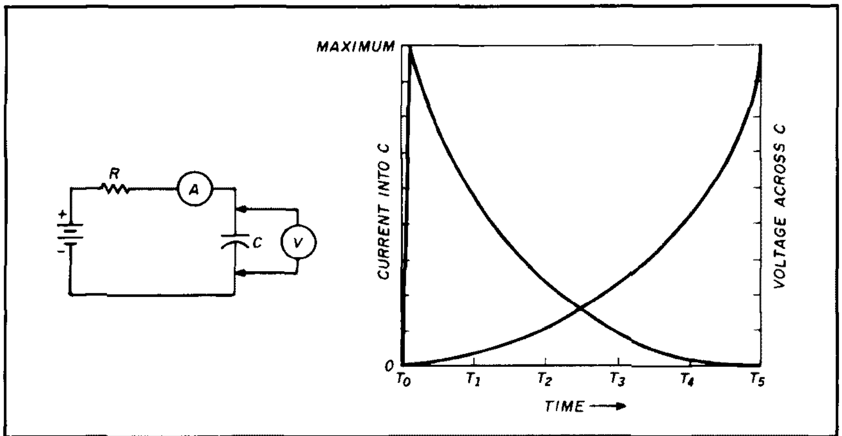
Do you remember the experiment I did with capacitors, resistors, and an LED? It required a definite amount of time for a capacitor to become fully charged when the voltage was applied through a resistor. The LED in the circuit glowed brightly at first, then slowly dimmed as the capacitor charged. This happened because the rush of current into the capacitor was large at first, then decreased as the capacitor became more fully charged. An "empty" capacitor has room for a lot of electrons, and a tremendous number of them rush to fill the space. As the capacitor fills up, the difference in potential (voltage) between the power supply and the capacitor drops and the electron flow decreases.

Figure 1 shows how this works. When the battery is first connected to the capacitor, the difference between the voltage across the battery and the voltage across the capacitor is very large, and the current rush is high. This is shown by the large peak near the left side of the graph. As time progresses and the voltage across the capacitor builds up, the current flow decreases. (There's a formula for determining the rate of charge and discharge of a capacitor. It's used to calculate the time factor in various circuits. I'll explore how to use this formula in another column.)

Figure 2 shows the voltage and current relationship in the circuit just discussed. The voltage is behind the current by 90 degrees (or the current *leads* the voltage by 90 degrees — whichever way you want to express it).

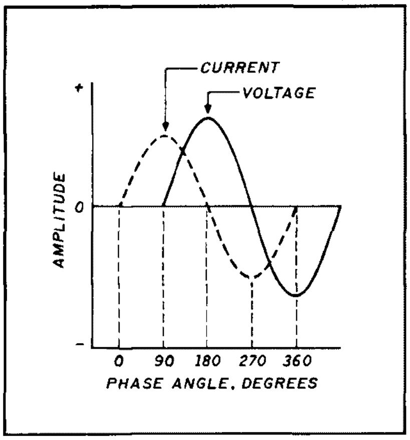
This characteristic of capacitors has a profound effect on alternating current. The power source tries to charge the capacitor each time the voltage starts to build up. But by the time the voltage has reached a peak, the current has already stopped. When the second half of the cycle starts, the current and voltage are again out of step. This is obviously a very inefficient circuit; maximum work isn't being done when maximum pressure (voltage) exists! In fact, if you pick the instant where the voltage is maximum and check the current flow, the power being used by the circuit is apparently zero. (Watts = Voltage × Current.) Actually, it isn't often that simple. The capacitor has stored some voltage, and when the AC voltage is at or near zero, the capacitor tries to release the

FIGURE 1



Charging current into a capacitor, and the voltage buildup across it, over a short period of time. Note that neither curve is linear.

FIGURE 2



In an alternating current (AC) circuit, a capacitor causes the voltage to lag the current, although convention generally states that the current leads the voltage. Compare this to the phase difference shown for inductance in last month's column where the current lags the voltage.

stored voltage back into the circuit, allowing some current to flow.

The difference between what's really happening (actual power) and what seems to happen (apparent power) determines the efficiency of an AC circuit, and is covered by the term *power factor*. A circuit with a poor power factor isn't working very well (more about this in a future issue). The power factor is important in AC circuits like the circuit that feeds 110 volts AC to your house. A related phenomenon is very important in Amateur Radio circuits like antennas and transmission lines, where a poor power factor causes something called standing wave ratio, or SWR.

How do we use this leading current?

This voltage/current difference can be used to correct the phase shift caused by an inductance. Suppose you have an audio amplifier with an inductance for frequency-response purposes. The inductor will cause the current to lag the voltage by some number of degrees. Placing the correct capacitor value in the circuit to make the voltage lag the current by the same number of degrees will put the voltage and current back in step again.

You could also employ the voltage/current relationship to obtain 180 degrees of phase difference. (An early "phasing" type of single sideband generator used this trick.) With a pair of amplifiers, one with an inductor to provide a +90 degree shift and another with a capacitor to provide a −90 degree shift, the difference becomes 180 degrees (see **Figure 3**).

The formula

Of course, there must be a way to calculate the effect of capacitance and the phase shift it introduces. A formula for this has evolved, just as it has with inductance. The term for this effect is *capacitive reactance*; the symbol for it is Xc. Two other familiar symbols are also involved — π and frequency (f). The formula is:

$$Xc = 1/(2\pi fC)$$

where f = frequency in hertz, and
C = capacitance in Farads.

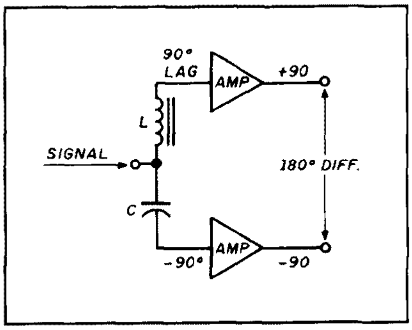
These units are rather large for Amateur Radio uses where you're often working with radio frequency circuits. It's more convenient to use smaller units for frequency, capacitance, and inductance. Here's a trick to remember. If you use MHz, μF, and μH together, you'll have fewer decimal places to contend with, and the answer will still be correct.

Now that you have this formula, you can predict the effect that a capacitor will have in a circuit. Let's say that you have a 0.00005-μF (50 pF) capacitor in a circuit working at a frequency of 3.5 MHz. Using the formula,

$$Xc = 1/(2\pi fC) = 1/(6.28 \times 3.5 \times 0.00005) = 1/0.00109 = 917.4 \text{ ohms.}$$

If this capacitor is placed across a 3.5-MHz signal, it will load it slightly. If it's placed in series, it will decrease the RF voltage in the same way a 917-ohm resistor would decrease voltage in a DC circuit.

FIGURE 3



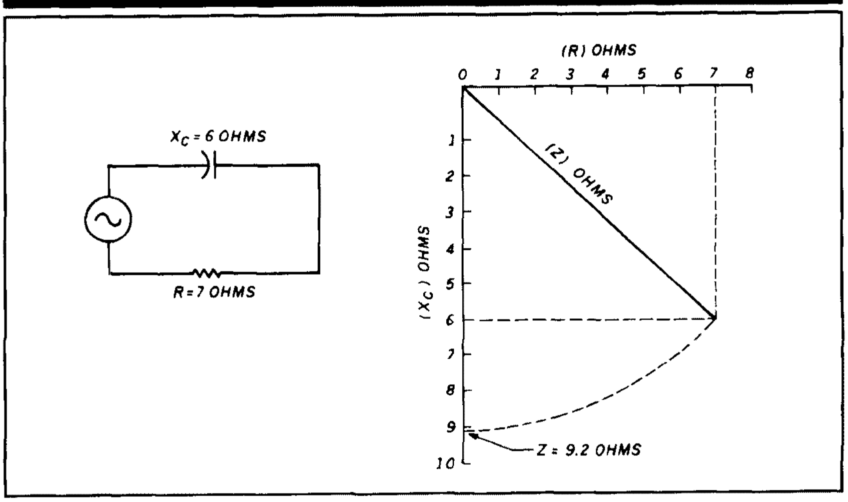
One use for both inductive and capacitive phase shift is to obtain a wide difference in phase, like the 180 degree difference at the output of these amplifiers.

Vectors, Act II

You thought you were through with vectors, didn't you? No such luck. Vectors are still with us, but you should find working with them easier this time around. You'll put them to work in more ways than one. All real-world circuits contain resistance in addition to capacitance. The vector diagram allows you to predict the effect of this combination. Impedance, Z, is again the symbol of interest, and can be obtained from a vector diagram as in **Figure 4**. The resistance is plotted horizontally; the reactance (Xc) is plotted vertically. Notice the difference between this vector and the one used for inductance in the last column. Both the capacitive reactance and its impedance vector are *below* the horizontal line. This is done by convention. It indicates that the capacitive reactance is negative (--), while the inductive reactance is considered positive (+).

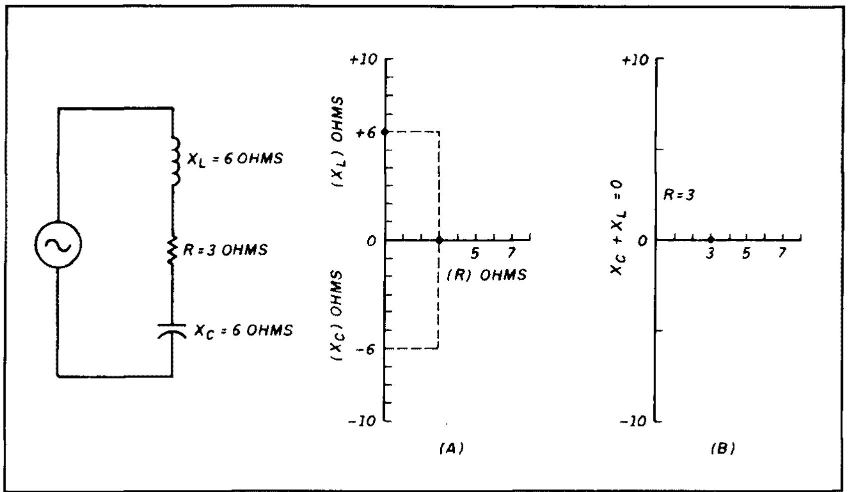
Here's a little secret. When you look at some of the formulas in many electronics textbooks, you often see a lower case letter "j" with a minus sign in front of it (−j). It usually has a number associated with it: −j3.8, for instance. Engineers call this valuable tool the j operator. This −j is telling you that there's some reactance in the circuit, and that the reactance is capacitive (the current is leading the voltage). Sometimes the −j value will also have a units designator, like −j58 ohms for a capacitive reactance value, or −j2.3 A to show that a reactive (leading) cur-

FIGURE 4



The vector diagram for capacitive reactance (Xc) and impedance (Z) is similar to that for inductive reactance, except that it is plotted below the horizontal resistance line.

FIGURE 5



Inductive reactances and capacitive reactances cancel when added vectorially, as shown here. An XL of +6 ohms and Xc of −6 ohms, with a resistance of 3 ohms plotted at (A), produces the vector diagram at (B) which shows only the 3-ohm resistance remaining. Such a circuit would appear to an AC signal as a 3-ohm resistor.

rent is flowing. You'll see a lot of these j operators in the design of transistorized circuits because most elements in a transistor have the same effect as placing a large capacitor across the signal.

It's important to know the effect of this capacitance so you can provide an inductance to counteract it; otherwise maximum power can't be transferred between stages. Some people place a + sign in front of the j when the reactance is positive (or inductive), and others don't.

Now let's get back to the vectors. To illustrate how the inductive and capacitive reactances cancel in a circuit, (A) in **Figure 5** shows a vector diagram of an inductive reactance and one of a capacitive reactance, plotted along the same vertical line. The circuit also contains a resistor, which is plotted horizontally to the right. The plus (+) and minus (−) vectors cancel; the result of combining the two is shown in (B) of **Figure 5**. All that's left is the simple resistance, R.

How many degrees?

There are a couple of ways to find the phase angle of a reactance. One method involves measuring the angle between the impedance vector (Z) and the resistance line (R) as shown in **Figure 4**. You may often do this, even when calculating the angle by other methods, just to see if you're on the right track. (A misplaced decimal point can really mess things up!) To get an idea of the magnitude of the angle, draw a horizontal line from the reactive (Xc) 7-ohm point and extend the resistive 7-ohm point to meet it. The line drawn from the 0 point to the intersection of these two 7-ohm lines is 45 degrees. The Z vector is less than this 45-degree line, so you know approximately where your answer should be if you use the second method correctly. The second method is to use the formula:

$$\text{Phase angle } \theta = COS{-}1(R/Z)$$

which is:

$$\theta = COS{-}1(7/9.2) = COS{-}1(0.7608) = 40.46 \text{ degrees.}$$

You can also use the SIN-1 of X/Z, or the TAN-1 of X/R to obtain the phase angle. Phase angle, by the way, is designated by the Greek letter theta (θ). Most pocket calculators have these trig functions; if yours doesn't, you can find trig tables in most math textbooks. However, the vector diagram will suffice for rough approximations.

Conclusion

Capacitive reactance in an AC circuit is just as much a fact of life as the inductive reactance covered in the last column. Fortunately, both can be calculated and used to obtain the desired effect needed when working with AC power, audio, or RF circuits. Some familiar uses in the Amateur Radio world include matching between stages in transmitters and receivers, getting correct phase shift for push-pull or parallel amplifiers, matching an output stage to a load, compensating for an antenna that's too long or too short, and (in digital circuits) delaying pulses in one part of a circuit to allow those in another part to catch up.

In my next column, I'll look at some more effects of combining XL and Xc, at the power factor and what it does to a circuit, and clean up a few loose ends associated with alternating current circuitry. hr

Practically Speaking

Joseph J. Carr, K4IPV

POTPOURRI

This month I'm going to talk about servicing radio equipment, especially older equipment. In the last year or so, I've had to service some elderly vacuum tube receivers as I added to my collection of antique ham and consumer radio receivers. Although these techniques are especially suited to older equipment, they will also work for many types of modern equipment. For example, you'll find the multi-tap resistor (which I'll discuss later) in antique gear and in certain relatively recent solid-state (except for the finals) imported HF transceivers.

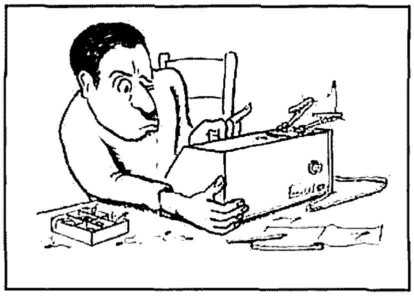
Tuning capacitor problems

Receiver, VFO, and transmitter RF problems are sometimes hard to diagnose. These problems can be so hairy that many Amateurs would rather not even bother with them. They are often very difficult to isolate and identify. That's why I was a bit disquieted when I first examined a Hallicrafters Model SX-100 general coverage shortwave receiver (the kind with the ham bands on an expanded bandspread dial) which I purchased at the Manassas, Virginia hamfest last year. I added this model to my collection of antique gear.

The big problems were in the main tuning capacitor. The capacitor problem manifested itself in several ways. The tuning was scratchy; it was loud; there were tunable oscillations; there were occasional abrupt, large changes of frequency while tuning; and the receiver would suddenly go dead, except for a little scratching while tuning.

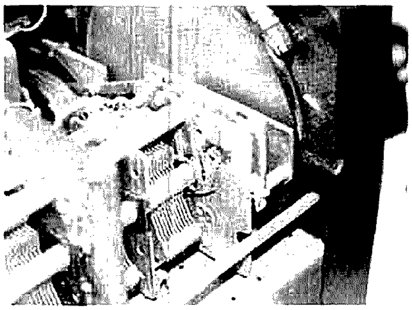
Photo A shows the receiver's multi-section main tuning capacitor. The rotor plates are grounded to the chassis because they are electrically connected to the capacitor's own mounting plate. The electrical connection of the rotor to the frame is made using one or more brass or steel spring "U" or "finger" clips. These clips straddle the rotor shaft at the mounting plate (located on the rear face of the front mounting plate in Photo A).

I found this clip was no longer making good electrical contact between the rotor shaft and the mounting plate. Corrosion had built up around the



spring clip contact points and under the rivet (see Figure 1). It was simple to effect a repair; I cleaned the corrosion from beneath the clip by slipping a relay burnishing tool beneath the end of the clip and the mounting plate. This tool is made of very thin pieces of spring steel and looks very much like a feeler gauge.

PHOTO A



Main tuning capacitor on SX-100. Gum built up under spring clip caused oscillations and other problems.

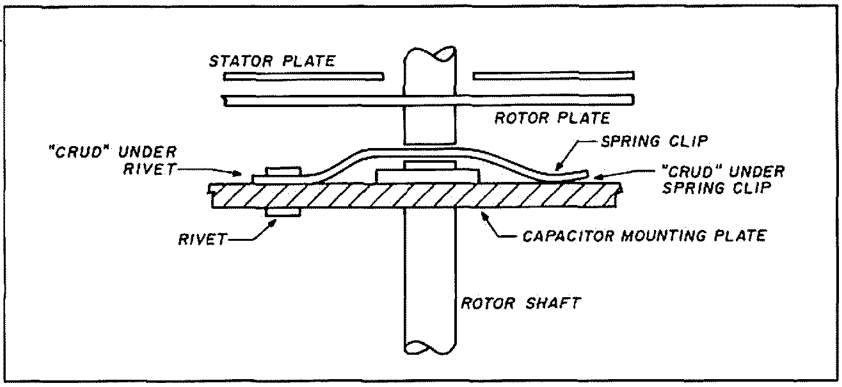
Several years ago I bought a Hammarlund HQ-145 from a life-long friend who had purchased it new in 1960, but never stayed with either ham radio or shortwave listening. It had been stored unused for nearly 15 years. The HQ-145 exhibited exactly the same problem as the SX-100 and also had some other problems common to older gear.

Other variable capacitor problems

Dried bearing lubricant. The lubricant in the ballbearing race on the front mounting plate of the capacitor often dries out. You can use ordinary aerosol switch contact cleaner (or liquid alcohol) to dissolve the old lubricant. But don't press the button on the aerosol can too hard. You want to make a quick, delicate spritz to keep from spraying fluid all over the capacitor or between the plates. Use a cotton swab to clean out the mess of old lubricant and cleaning fluid. Once you've cleaned the bearing race, refill it with a dab of white lubricant, like Lubriplate®, using a toothpick as an applicator. Be careful to keep lubricant from getting between the plates.

Shorted Plates. There are two things that cause short circuits between the plates of the variable capacitor. One is foreign matter (including dust or metal particles) that collects inside the plate assembly; the other is bent plates. You can often dislodge the foreign matter with a quick blast or two of dried compressed air. Electronics parts and supplies stores, some auto parts stores, and photo supply stores sell small cans (similar to aerosol cans) of dried compressed air just for this purpose. You usually have to buy a

FIGURE 1



Detail of the grounding spring clip on the SX-100 capacitor.

nozzle attachment for the can, although I've seen one type with a plastic nozzle fitted to it.

Bent plates are another matter. If the bent plate is close to the surface, small needle nose pliers will be useful for rebending the plate to its original shape. Otherwise, you may have to use a small tool like the burnishing tool to work the plate gently into the correct shape and position. *Do not* use a file or other cutting tool. These tools leave filings that will short the plates together even more.

Some people "align" receivers and VFOs by bending the plates of the tuning capacitor; this isn't very smart. The original design was probably good enough to allow proper dial tracking using the right balance of main capacitor, trimmer capacitors, padder capacitors, and inductor settings.

Shorted capacitor. Sometimes you'll hear a scratching sound as the tuning capacitor is tuned across the band. This is a good indication that you have a short in the capacitor. However, the scratching could also be caused by the spring clip used to ground the rotor. One way to tell whether or not the capacitor is shorted at one point, and where that point is in the rotation of the capacitor, is to use an ohmmeter. Although you can use both analog and digital ohmmeters, analog meters are a little easier to work with in this application.

Set the ohmmeter to its highest resistance scale. Disconnect the capacitor from the circuit, and connect it across the probes of the ohmmeter. Tune the capacitor *very slowly* through its entire range while watching the ohmmeter. If there's a positional short, the ohmmeter will flick downscale when you've located it. You may need a good eye and a strong magnifying glass to see which of the many pairs of plates are actually shorted, but the problem should be visible.

Scummed, gummed, and gunked capacitors. Variable capacitors are natural depositories for all manner of nasty stuff — especially if the radio was used in a greasy environment. Airborne grease settles on the capacitor's plates; then dust settles on the grease. Cleaning up this type of mess is decidedly difficult. There are, however, several aerosol degreasers available from radio supply stores. You can also use a product like Birchwood Casey Gun Scrubber (available in gun shops). Firearms suffer a fate similar to that of capacitors because they are lubricated, collect dust, and (above all) fill up with the residue of spent and unspent gunpowder that showers the piece when it's fired.

WARNING! *Do not use carbon tetrachloride to clean the capacitor.* Many early radio books recommended this chemical as a cleaner, but "Carbon Tet" is now known to be a health hazard.

Making odd value capacitors and resistors

There's a list of standard resistor and capacitor values that meet a large number of needs. However, there are cases when a project you're building requires an oddball part value. "Standard" resistor and capacitor values have also changed over the years, so you may find that the rig you're repairing has a bad component of an odd value which is no longer available. Whenever possible, you should replace the component with one of identical ratings — but there are times when that isn't possible.

Sometimes, a close standard value can be substituted for a "once" standard value without a deterioration of radio performance. For example, I had a receiver with a 16-k, 1-watt screen dropping resistor, and replaced it with a 15-k, 1-watt resistor with no problem. Remember, however, that the physical size of the new resistors is smaller than the older ones, especially among the carbon composition replacements (metal film, flame proof) used today.

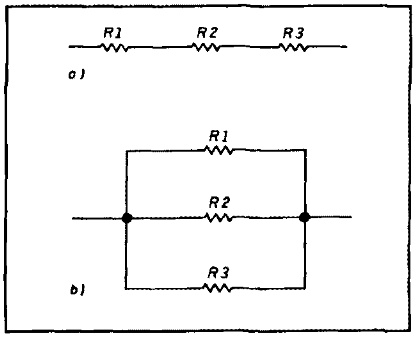
Figure 2 shows two ways to obtain odd values of resistance. In **Figure 2A**, two or more resistors are connected in series. The total resistance is the sum of all the individual resistors or, in the case of the three resistors shown, $R_t = R1 + R2 + R3$. The total wattage rating is the sum of the individual wattage ratings, if all three resistors are equal. (They would have equal voltage drops.) If you have unequal resistances, you must calculate either the voltage drops across each resistance or the total current through the resistors to find the power dissipation of each one, and prevent overloading one particular resistor. The following equations apply:

$$P = (V_d)^2/R \quad \textbf{(1)}$$

$$P = I^2/R \quad \textbf{(2)}$$

where:

FIGURE 2



(A) Resistors in series, (B) resistors in parallel.

P is the power dissipation in the resistor
R is the resistance of the resistor
I is the current flowing in the resistor
V_d is the voltage drop across the resistor

To calculate the voltage drop across any one resistor, you must find the proportion of the whole that particular resistance represents. With n resistors in series, the voltage drop across the nth resistor is:

$$V_{di} = V \times \frac{R_i}{R1 + R2 + R3 + ... + R_n} \quad \textbf{(3)}$$

where:

V is the applied voltage across the entire series chain of resistors
V_{di} is the voltage drop across the nth resistor
R_i is the resistance of the nth resistor
R1, R2, R3, and R_n are the individual resistances of all n resistors in the series circuit

Figure 2B shows a parallel combination of resistors. The total resistance of a parallel combination is always less than the resistance of the lowest value resistor in the network. For all parallel resistor networks with n resistors:

$$R_t = \frac{1}{\frac{1}{R1} + \frac{1}{R2} + \frac{1}{R3} + ... + \frac{1}{R_n}} \quad \textbf{(4)}$$

In the special case of two resistors in parallel, you can use either **Equation 4** or the modified version below:

$$R_t = \frac{R1R2}{R1 + R2} \quad \textbf{(5)}$$

If the network is made up of identical resistors (all resistors having the same value), the total resistance is the value

of any one of them divided by the number of resistors in the network. Thus, three 100-ohm resistors in parallel have a total resistance of 100 ohms/3, or 33.33 ohms.

The total wattage rating of the parallel network is the sum of the individual wattage ratings — assuming that the resistors are equal, or very nearly so. Otherwise, you'll have to calculate the current in each resistor using Ohm's law (I = V/R) and then the power dissipation, in order to ensure that the wattage rating of any one resistor isn't exceeded.

Capacitors can also be combined to form other capacitance values, although somewhat less successfully in some circuits. This is because things like distributed or stray values of capacitance and inductance (yes, practical capacitors have an inherent inductance because of their construction) are sometimes unpredictable.

Figure 3A shows capacitors in parallel. The total capacitance of this network is the combination of like resistors in series; that is, it's the sum of the individual capacitances. The total working voltage of the combination is the lowest working voltage rating of all the capacitors. In other words, if 100 working volts DC (WVDC), 1,000 WVDC, and 600 WVDC capacitors are connected in parallel, then the WVDC rating of the combination is 100 WVDC.

Capacitors in series combine like resistors in parallel. You can calculate the total capacitance of a series capacitor network with the same equations used for parallel resistors, but substitute C1, C2, etc. for R1, R2, and so on.

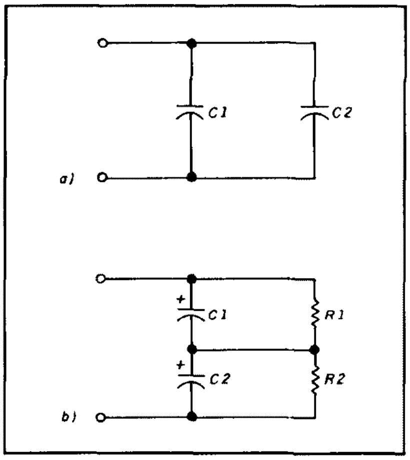
If the capacitances are equal, the WVDC rating of the series combination is the sum of all individual WVDC ratings. Otherwise, the voltages combine in a manner that reflects the capacitance of each unit. This can be a problem if there's an AC component or (in the case of power supply circuits) a pulsating DC ripple riding on the DC level.

There's a special solution for balancing the voltage drop across series electrolytic capacitors (see **Figure 3B**). Connect a resistor in parallel with each capacitor. The value of this resistor is usually about 100 ohms per volt of the WVDC rating, roughly 27 to 470 k in common circuits. All resistors in the series stack should have the same value. The power rating of the resistors should reflect a good safety margin over the actual dissipation.

Is that transformer shorted?

The power transformer in a receiver or low power transceiver can be a little difficult to "ring out" using just an ohmmeter, but an open winding is obvious and shows up easily. In the case of a massive short, the transformer will grossly overheat. It may smoke, ooze tar, and will probably blow the fuse. But what about an "in between" case where a winding is partially shorted? **Figure 4** shows a method for testing the radio transformer. First, disconnect all of the secondary windings or remove all the tubes from the radio to reduce current drain. Next, connect the radio AC line cord to an outlet box which has a 25 to 40-watt, 115-volt AC lamp connected in series with one line of the outlet. Turn the radio on and apply AC power to the outlet box, noting the brightness of the lamp. A good transformer will barely glow if the transformer remains connected with the tubes removed; a bad transformer will cause the lamp to glow brightly.

FIGURE 3



(A) Capacitors in parallel, (B) capacitors in series with balancing resistors.

Replacing the volume, tone, or RF gain controls

The volume control, tone control, and RF gain (if used) controls on a radio are either potentiometers or, on the very oldest sets, rheostats. You can buy new controls at radio/TV parts wholesalers. However, you'll need to know a few things before buying. Ask these questions:

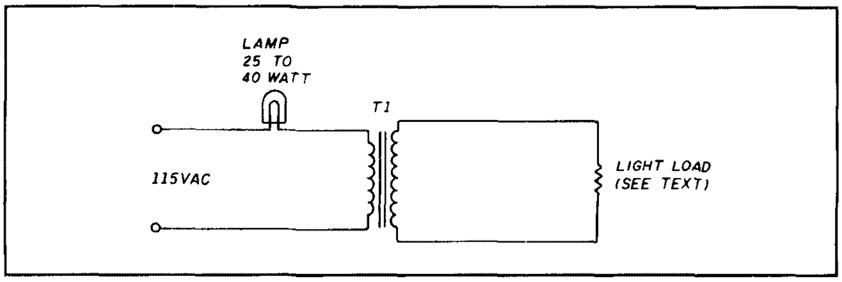
- What's the resistance? (Measure across the two outer terminals.)
- Is it wirewound or carbon? (Look at the element; most are carbon.)
- Is the control audio or linear taper? (Use audio for volume controls and linear for the others.)
- Does the volume control have a loudness tap? (Look on the underside for a "spare" terminal.)
- Is the shaft half round or full round? Is it metal or an insulated material? Is the shaft smooth or "splined?"
- Is there a switch mounted on the rear of the control? If so, how many terminals are there?

Once you've selected the control, you'll probably find that its shaft is too long (a consequence of "universality"). **Figure 5** shows how to measure the shaft of the new control against the old one. Mark the shaft of the new control at the point indicated in the figure using a hacksaw blade or scribe. Next, place the control in a bench vise and cut the shaft off at the mark.

Replacing multi-tap resistors

Many receivers have multiple resistor packages containing two or more series resistors. These resistors are basically one long resistance element tapped at the desired points. They are often shown in schematics as separate resistors, although some appear as an

FIGURE 4



Test circuit for finding a shorted transformer.

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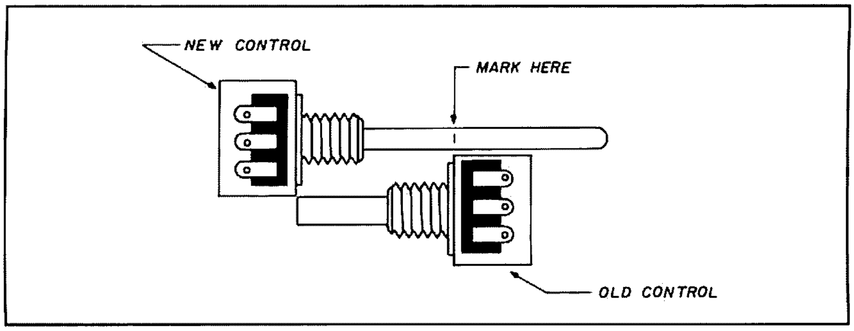
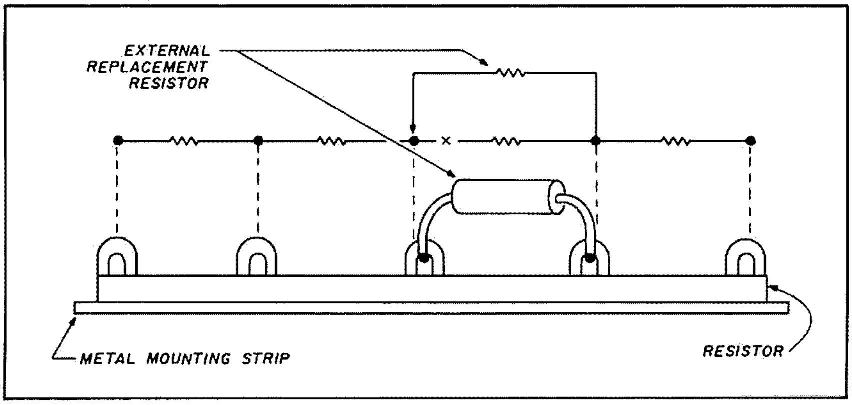
FIGURE 5



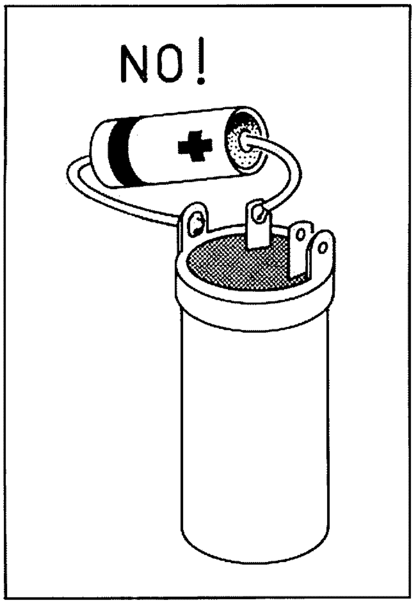
Measuring replacement potentiometer shafts.

FIGURE 6



Replacing a bad section of a multi-tap resistor.

FIGURE 7



***Bad practice!* Don't permanently bridge a good electrolytic capacitor across a bad section of a multisection capacitor. Replace the whole capacitor.**

assembly. When one section goes open, it can be bridged with a separate power resistor (see Figure 6). I've also shown two methods of mounting in Figure 6; select the one that seems reasonable at the time.

A bad practice!

Recently I went to visit a friend (not a ham — yet) who repairs and restores antique radio sets. He asked me, somewhat plaintively (knowing that I was a radio repairman "way back when" in 1959), if it was common in the olden days (thanks a lot!) to bridge a tubular electrolytic across an open section of a multisection electrolytic capacitor. Unfortunately, I had to tell him that this practice (see Figure 7) was all too common. But it's an extremely bad one and should be avoided whenever possible. The problem with this practice is that the original capacitor is still defective, and may later short out to ground — or to another section. So, in addition to knocking your receiver or transceiver off the air, the defective capacitor may also cause secondary damage to the rest of the power supply circuits.

SHORT CIRCUITS

Dear Readers:

Well the gremlins were certainly at work when we put the March issue together. Here are a few errors you spotted as you browsed through the pages! Ed.

New Product Correction

In the new product listing, "Two Rotatable Dipoles," on page 73 of the March 1990 issue, there were two errors. The antenna elements are made from high strength 6061-T6 aluminum alloy, not T-60612. Also the address given was incorrect. SV Products is located in Woodburn, Indiana — not Illinois.

For those of you who are interested, SV Products has just released a new two-element dual band, computer optimized antenna — the Model 1824 2L — for the 12/17 meter band. This antenna has a new trap design that gives high mechanical strength and power handling capability. It's available at the introductory price of $199.95. Contact Gary Nichols, SV Products, 4100 Fahlsing Road, Woodburn, Indiana 46797 for details.

K3OF Supercharger Boards Available; Missing Parts List Located

We neglected to mention that printed circuit boards are available for the project described in "Build Your Own Supercharger," by K3OF, March 1990, page 22. You can order them from FAR CIRCUITS, 18N640 Field Court, Dundee, Illinois 60118. The cost is $2 for the small board with a single charging circuit and $9.75 for the large board, which includes 8 charging circuits. Please include $1.50 shipping and handling with each order.

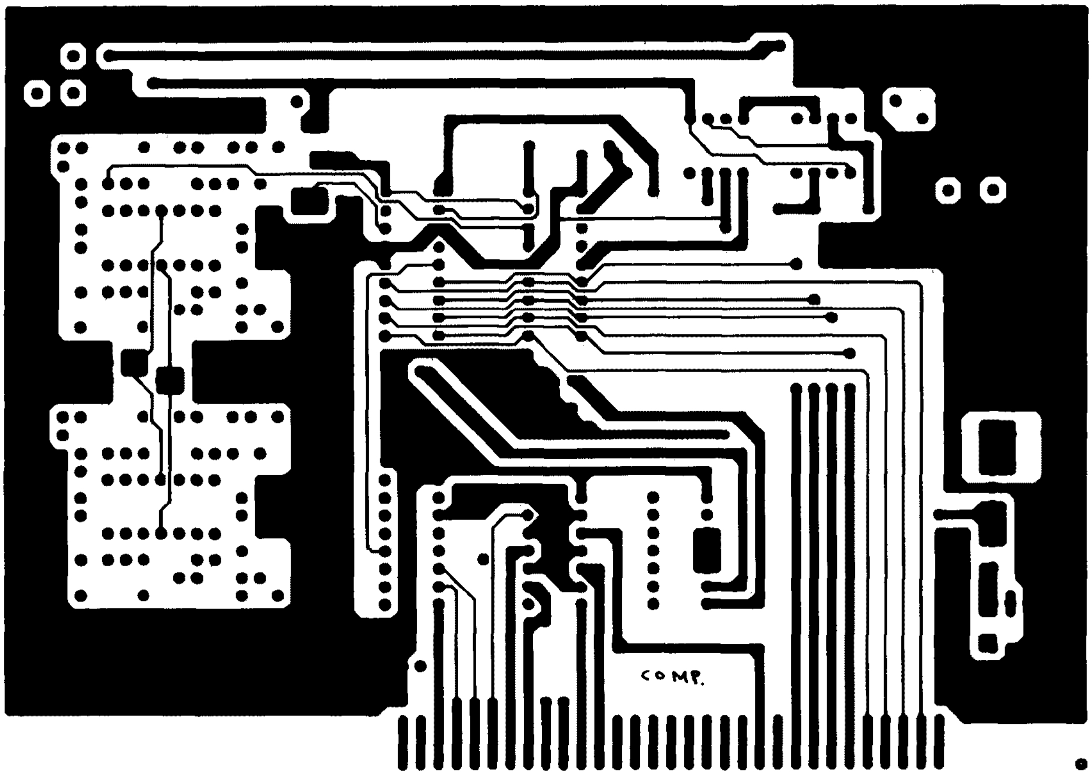
We also forgot to include the parts list for those of you who'd like to build this handy little gadget. Here it is!

PARTS LIST

		Source/Approximate Cost			
For each regulator circuit		*Jameco**	*Cost*	*Radio Shack*	*Cost*
U1	*LM723CN*		*$0.39*	*276-1740*	*$0.99*
Q1	*Tip-3055*		*0.69*	*276-2020*	*1.59*
	(or Tip31)		*(0.49)*	*276-2017*	*(0.99)*
C1	*100-500 pF*	*MD220/50*	*0.10*	*272-124*	*0.20*
R1,R3	*1 k*	*R1K*	*0.10*	*271-1321*	*0.16*
R2	*5 k potentiometer*	*63P5K*	*0.89*	*271-217*	*0.69*
R4	*Value as required***		*0.05*		*0.08*
14-pin dip socket		*14LP*	*0.12*	*276-1999*	*0.45*
	Cost per basic circuit		*$2.34*		*$4.18*
Miscellaneous and optional parts:					
R5	*20 ohm*	*R20*	*$0.05*		
R6	*750 ohm 1/2 watt*				
S1	*SPDT switch*	*MPC123*	*0.99*	*275-635*	*$2.89*
CR1	*Full wave bridge rectifier, 4 A*			*276-1146*	*1.39*
C2	*2200 µF, 35 volts*			*272-1020*	*2.49*
Heat sink TO220		*291-36H*	*0.39*	*276-1363*	*0.79*
LED as desired					
PC board as required					
Project box as desired					

Supply transformer: Any 18 to 24 volts AC or DC 0.5 A suggested.
All resistors may be 1/4 watt except R6, which is 1/2 watt.

**Jameco Electronics, 1355 Shoreway Road, Belmont, California 94002. $20 minimum order. Minimum quantity of 10 of each value resistor.*
***Two 10-ohm resistors may be paralleled for 5 ohms if a 4.7-ohm resistor isn't available. Radio Shack doesn't stock resistors below 10 ohms. Two 1.5-k 1/4-watt resistors may be paralleled for the 750-ohm 1/2-watt resistor (R6).*



Double-sided Board Trouble

Those double-sided pc boards can be tricky. We gave you one side of WA4ADG's board in his March 1990 article "Digital Voice Storage in the Ham Shack," (page 56) and showed you where the components should be placed on the component side of the board, but we didn't print a reproducible copy of the side of the board. You'll find it above. There are also two changes in resistor values. Resistor R1 (**Figure** 1, page 61) should be 51 k, not 5.1 k. The resistor connected to pin 3 of IC1 (**Figure** 1, page 61) should be 18 k, not 36 k.

DX Forecaster

By Garth Stonehocker, K0RYW

1989 DX Propagation Conditions

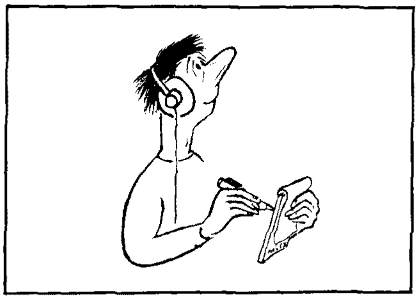
DX propagation conditions for 1989 were characterized by high maximum usable frequencies (MUFs) and many large solar flares with related disturbances of the geomagnetic field. This is typical during a near sunspot maximum year, when the solar flux average for that year is 214. The solar flux was high (237) in January and then decreased each month until April and May, when it dropped to 190±1. June bounced up to 242 with the highest daily maximum (326) thus far in sunspot cycle 22. July solar flux dropped to the year's monthly minimum of 183. The rest of the year showed a slight upward trend from August (218) to November (234); October and December were a bit lower.

For propagation and research purposes, solar activity is tracked in 27-day sun rotation cycles. The low months during the year lacked very high monthly 27-day maximums when compared with the high activity months. This effect is typical for the solar flare activity and flux which arrives at the earth as the sun rotates and brings its active regions into view.

Some propagation details

We define ionospheric conditions by vertically measuring the height (time of pulse travel) of the ionosphere at various frequencies (even continuously) to give the maximum frequency (foF2), or measuring the MUF for an oblique path between a pulse transmitter and its receiver. In creating **Table 1**, I looked at mid-latitude foF2s (MUFs are about three times foF2) for the months by using the highest diurnal median (the value available 50 percent of the days of the month) frequency for each month. I've also given the highest peak value of daily foF2 at any daytime hour during the month. These foF2s are displayed in a month-by-month summary for the year.

As you look over these numbers, compare the difference between peak and median foF2s. A comparison of the foF2s with the month-to-month



direction of the solar flux values shows an interesting effect. The ionosphere's foF2 is generated by an upward drift of ionization to around 250 to 350 km from its 80 to 180-km production altitude. The lower production altitude varies much more closely with solar flux ultraviolet than foF2 height. You'll note a few days delay in peak foF2 at a 27-day solar flux peak and a general smoothing across foF2 monthly averages for the real "median" values. The July minimum, which occurred between the two higher months of June and August, didn't decrease the foF2 median much — but geomagnetic disturbances can affect foF2 median much more quickly and to a greater degree. Remember that the decrease in mid-latitude foF2 resulting from geomagnetic disturbance is really the auroral zone (an oval of incoming solar wind particles) moving equatorward while pushing the ionospheric trough (low foF2) before it. The trough lost more of its ions. They moved up the magnetic field lines, ending up at 20 degrees north or south of the geomagnetic equator. This gave us the one long hop transequatorial propagation in late evenings in the Northern Hemisphere during the winter and equinoxial months. During disturbances mid-latitude foF2s go down and equatorial foF2s go up. You can see the mid-latitude foF2 decreases in some of the differences of the peak to median columns in **Table 1** as compared with the number and highest geomagnetic "A" values column for the month. Even months with a low flux have high foF2s — if there have been few disturbances. When raised by the 27-day SSN flux, the high SSN basic foF2 increases absorption in the D and E region to the point that the signal's strength decreases — even up to 10 meters. If we had Amateur bands between 6 and 10 meters, strengths would show increases near the MUF up there. Unless the MUF increases to 6 meters, there is no chance to witness this. You can, however, see an increase in the number of hours 10 meters is open. Enjoy the high MUFs or long openings and disturbance-related DX from unusual locations while you can. We still have a couple of years to go in this sunspot cycle. But the increase in solar flares and increased level of disturbances hold down MUF by as much as 20 to 40 percent on many days — down into the 10-meter range again.

Last minute forecast

The first week of the month favors nighttime openings on the lower frequency bands. Then the solar flux is expected to build up, causing the MUF to increase and lengthening time for openings for the higher frequency

TABLE 1

Month by month foF2 summary for 1989.

Month	foF2, MHz Median	Peak	Number and Highest "A"	Average Flux
January	12.9	15.0	4/33	237
February	13.0	13.8	2/33	223
March	12.6	13.8	5/248	207
April	10.4	13.2	2/38	189
May	8.2	11.4	2/40	192
June	7.2	10.6	3/49	242
July	7.7	10.5	0	183
August	8.6	10.2	4/61	218
September	10.7	11.9	3/55	226
October	13.2	14.2	1/86	207
November	14.2	15.5	3/71	234
December	13.4	15.3	5/36	213

MAY

WESTERN USA

GMT	PDT	N (Asia Far East)	NE (Europe)	E (S. Africa)	SE (S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania Australia)	NW (Japan)
0000	5:00	12	_17_	15	10	10	10	10*	10
0100	6:00	12	_20_	_15_	10	10	10*	10*	10
0200	7:00	10	_20_	_15_	10	12	10*	10*	10
0300	8:00	12	_20_	_17_	10	15	10*	10*	10
0400	9:00	12	17	_17_	12	17	10	10	10
0500	10:00	12	15	15	15	20	10	10	10
0600	11:00	15	15	10	15	20	10	10	12
0700	12:00	15	17	10	_15_	20	10	10	12
0800	1:00	17	17	12	_17_	20	10	10	15
0900	2:00	_17_	_20_*	15	_17_	20	10	12	15
1000	3:00	_20_	15	15	_17_	20	12	15	17
1100	4:00	_17_	15	15	_20_	20	15	15	17
1200	5:00	15	12	10	15	_20_	15	15	_30_
1300	6:00	15	12	10	12	_20_	_15_	17	_20_
1400	7:00	17	12	10	10	_20_	_15_	_17_	_17_
1500	8:00	15	10	10	10	_20_	_17_	_17_	_17_
1600	9:00	15	10	10	10	_20_	_17_	_17_	_20_
1700	10:00	17	10	10	10	15	12	_17_	_20_
1800	11:00	_17_	10	10	10	12	10	_17_	17
1900	12:00	_20_	12	10	10	12	10	12	15
2000	1:00	17	12	10	10	10	10	10	15
2100	2:00	15	12	10	10	10	10	10	12
2200	3:00	15	15	12	10*	10	10	10	12
2300	4:00	12	17	12	10	10	10	10	10

MID USA

GMT	MDT	N (Asia Far East)	NE (Europe)	E (S. Africa)	SE (S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania Australia)	NW (Japan)	CDT
0000	6:00	12	15	15	10	10	10*	10*	10	7:00
0100	7:00	12	17	_15_	10	12	10*	10*	10	8:00
0200	8:00	12	_17_	_17_	10	15	10	10*	10	9:00
0300	9:00	12	_20_	_17_	10	15	10	10	12	10:00
0400	10:00	15	_30_	15	12	17	10	10	12	11:00
0500	11:00	15	_20_	15	15	20	10	10	12	12:00
0600	12:00	17	17	_15_	15	20	10	10	15	1:00
0700	1:00	_17_	15	15	_15_	20	10	10	15	2:00
0800	2:00	_20_	17	15	_15_	20	12	12	17	3:00
0900	3:00	_17_	_20_	15	_17_	20	12	12	17	4:00
1000	4:00	15	15	12	_17_	20	15	15	_20_	5:00
1100	5:00	15	15	10	_17_	20	15	15	_17_	6:00
1200	6:00	17	12	10	15	_20_	17	17	_17_	7:00
1300	7:00	15	12	10	12	_20_	17	17	_20_	8:00
1400	8:00	15	12	10	10	_20_	_15_	_17_	15	9:00
1500	9:00	12	10	10	10	_20_	_15_	_17_	15	10:00
1600	10:00	12	10	10	10	_20_	_17_	_17_	_17_	11:00
1700	11:00	15	10	10	10	17	_17_	_17_	_20_	12:00
1800	12:00	17	12	10	10	15	15	_17_	17	1:00
1900	1:00	_17_	12	10	10	12	12*	12	15	2:00
2000	2:00	_30_	12	10	10	12	10	10	15	3:00
2100	3:00	_17_	15	10	10	10	10	10	12	4:00
2200	4:00	15	15	12	10	10	10	10	12*	5:00
2300	5:00	15	15	12	10	10	10*	10	10	6:00

EASTERN USA

GMT	EDT	N (Asia Far East)	NE (Europe)	E (S. Africa)	SE (Caribbean S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania Australia)	NW (Japan)
0000	8:00	12	15	15	10	12	10*	10*	12
0100	9:00	12	17	15	10	15	10*	10*	12
0200	10:00	15	17	_15_	10	15	10	10	12
0300	11:00	15	_30_	_17_	10	20	10	10	12
0400	12:00	17	_20_	_17_	12	20	10	10	15
0500	1:00	_17_	_20_	15	15	20	10	10	15
0600	2:00	_20_	_30_	15	_15_	20	10	10	17
0700	3:00	17	_20_	15	_15_	20	12	12	_30_
0800	4:00	15	17	_15_	_17_	20	12	12	_17_
0900	5:00	15	15	12	_17_	20	15	15	15
1000	6:00	12	15	10	_15_	20	15	15	15
1100	7:00	12	12	10	12	_20_	15	15	12
1200	8:00	12*	12	10	10	_20_	17	17	15
1300	9:00	12*	12	10	10	_20_	_15_	17	15
1400	10:00	12	12	10	10	_20_	_15_	_17_	17
1500	11:00	12	10	10	10	_30_*	_17_	_17_	_17_
1600	12:00	15	10	10	10	17	_15_	_17_	_20_
1700	1:00	15	10	10	10	15	12	_17_	_17_
1800	2:00	17	12	10	10	12	10	_17_	15
1900	3:00	_17_	12	10	10	12	10	10	15
2000	4:00	_30_	12	10	10	12	10	10	12
2100	5:00	_17_	12	10	10	10	10	10	12
2200	6:00	15	15	12	10	12	10*	10	12
2300	7:00	15	15	12	10	12	10*	10*	10

✓ 175

✓ 176

bands into the second and third weeks of May. But these weeks will include signals weakened by absorption and a chance of disturbance, with fading and fluttery signals around the 9th, 15th, and 24th. Paths to the south may have short 6-meter openings. Nighttime DX paths may have better openings during the fourth and last weeks of the month. These paths are expected to be the most affected by the disturbances, but you may receive DX surprises from new countries.

The full moon occurs on the 9th; the lunar perigee is on the 26th. The Aquarid meteor shower (for meteor scatter and meteor burst DXers) peaks between May 4th and 6th, with rates of 10 and 25 per hour for the Northern and Southern Hemispheres, respectively.

If you want more help with your DXing, try a DX net. The 1990 edition of OE2DYL's popular list "DX Nets Around the World" (List 9) is available. It contains data on more than 100 active DX nets. The price is $3, U.S. (airmail). The package price for editions 1 to 9 is $12, U.S. ($1 = 3 IRC. Sorry, no checks.) Send your order with an SASE to Dieter Konrad, Rosengasse 1, 5020 Salzburg, Austria.

Band-by-band summary

Six meters may have long skip openings to the southeast and southwest late afternoons on days when the solar flux is high. Ten, 12, 15, 17, and 20 meters will support DX propagation from most areas of the world during daylight hours and into the evening, with long skip out to 2000 miles (3500 km) per hop. Signals on the upper three bands will be strongest from the southern countries and occur near local noon. The propagation direction will follow the sun across the sky. Look to the east in the morning, the south at midday, and the west in the evening. Sporadic E short skip will be available at local noon on some days toward the end of the month.

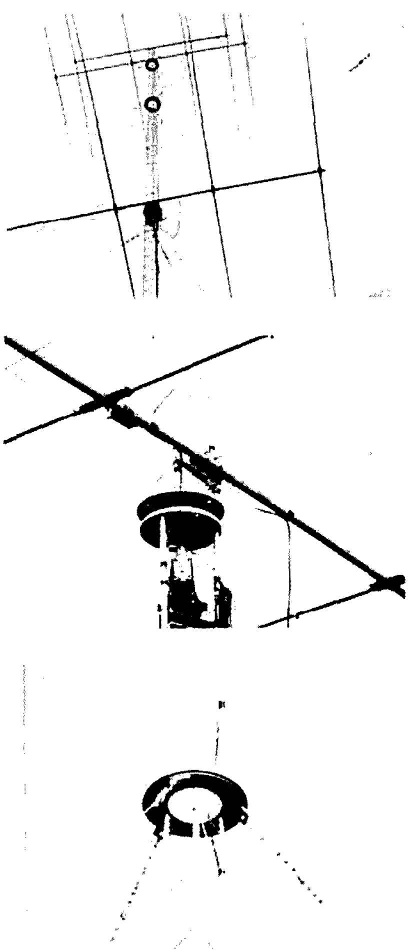
Thirty, 40, 80, and 160 meters are the nighttime DX bands. Propagation follows the darkness path across the sky. Look to the east in the evenings, to the north and south around midnight, and toward the west in the predawn hours. Skip distances will generally decrease to 1000 miles (1600 km) on these bands. Sporadic E openings will be most frequently observed around sunrise and sunset toward the end of the month. ham radio

NEW PRODUCTS

New Hardware for Rohn 25 Rotating Tower

Rotating Tower Systems, Inc. announces the availability of guy wire bearings and rotating base assemblies designed especially for use with Rohn 25 tower. This hardware lets you build a versatile and economical rotating tower using Rohn 25 towers sections and companion rotators. Rotating stacks of smaller monobanders, stacked tribanders, large VHF/UHF arrays, and antennas mounted at optimum heights are applications for a rotating tower made from these components. In addition, component design allows the rotating base unit to be mounted at any tower height minimizing the number of guy wire bearings. For information and prices write or call Rotating Tower Systems, Inc., Box 44, Prosper, Texas 75078. Phone: (214)347-2560.

Circle #305 on Reader Service Card.



Reintroducing the Heights Tower System

Heights Tower Systems reintroduces its "standard" aluminum Heights tower. These towers are once again being manufactured to their original specifications.

Heights has a complete line of regular tapered towers and 34 configurations of telescoping crank-up towers. All are built of non-corrosive, high-tensile strength aluminum. Recommended tower configurations range from heights of 89 feet with 520 pounds of windload capacity up to 120 feet with 160 pounds of windload. Tower sections measure 8 feet in length. The tower can be assembled completely on the ground.

Hinged bases and fold over kits are also available. Hinged bases let you raise or lower the tower, "hinged," on two of three base legs via specially drilled stubs. Fold over kits, with screw or winch-operated folding plates, allow one person to fold over the tower. Fold over kits may be operated manually or can be motorized for electrical operation. An optional 4-foot stand mounts the fold over kit at your standing level. The folding plates may be installed at 8, 16, or 24-foot high junctures in the tower while being controlled from below. Most tower accessories, including the fold over kits, can be fabricated in regular or stainless steel, or aluminum, at your request.

For more information, write or call Heights Tower Systems, 1721 Indian Road, Lapeer, Michigan 48446. Phone: (313)667-1700.

Circle #312 on Reader Service Card.

New Analog Trainer from Elenco

Elenco Electronics, Inc., has a new analog trainer. Model XK-120 is designed for students who are learning the fundamentals of analog circuits. Circuits are easily assembled on the large 840-pin breadblox work area. The trainer has two DC power supplies and one AC supply which are all regulated and protected against shorts. Also provided is a function generator capable of producing sine, square, or triangle waveforms up to 100 kHz.

The trainer comes in a carrying case with a lid compartment for holding experiment parts. It's available in kit form with easy-to-follow instructions and a troubleshooting guide. The cost of the assembled version of the XK-120 is $150; in kit form the price is $110.

For more information contact Elenco Electronics, Inc., 150 W. Carpenter Avenue, Wheeling, Illinois 60090.

Circle #302 on Reader Service Card.

ICOM's Mini-Handheld Communications Receiver

The ICOM IC-R1 continuously covers 100 kHz to 130 MHz with AM, FM, and wide-FM modes. Measuring just 1.9" × 4" × 1.4", this tiny receiver features:

- built-in NiCd batteries
- multi-function scanning
- keyboard and tuning
- 100 memory channels
- built-in clock

The IC-R1 is also equipped with a power saver function, adjustable LCD contrast, a signal indicator, an external DC power jack with battery charge capability plus a large variety of other options from the "S" series handhelds.

For details contact ICOM America, Inc., 2380 116th Avenue, NE, P.O. Box C-90029, Bellevue, Washington, 98009-9029.

Circle #303 on Reader Service Card.

ALPHA DELTA Model CLP Rotor Control Line Transi-Trap Surge Protector

The ALPHA DELTA Model CLP Rotor Control Line Transi-Trap Surge Protector protects your communications equipment against lightning-induced surge voltages on the control lines to your rotor control and remote antenna switch boxes.

The protector features straightforward installation with no soldering and:

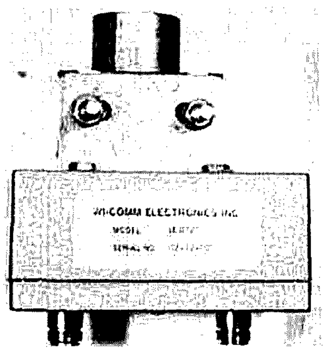
- Protects up to eight 16 AWG wire control line cables. Covers the most commonly used rotor and remote switch models. Requires no modification to control boxes.
- Uses eight NEMP-rated high surge current field-replaceable gas tube Arc-Plug cartridges. Each line is individually protected.
- Has quality G-10/FR4 glass epoxy pc board construction. Your control line connects directly to industrial grade pc board mount connectors for best low inductance discharge performance. Computer designed. No soldering required.
- Equally effective for modem/phone line protection. The low capacitance gas tube Arc-Plug cartridges accommodate high baud data transmission.

The Model CLP Control Line Transi-Trap Protector sells for $49.94, with $3 added for postage and handling. Transi-Trap coax cable surge protectors are also available. (Ohio residents add 6-1/2 percent sales tax.) Write Alpha Delta Communications, Inc., PO Box 571, Centerville, Ohio 45459 for more details. Phone: (513)435-4772.

Circle #307 on Reader Service Card.

New Tower-mount Preamplifier

WI-COMM Electronics, Inc. offers a new tower-mount preamplifier in a weatherproof aluminum housing which covers the band from 10 to 900 MHz. The preamplifier gain is 25 dB, noise figure is 3.1 dB typical, 1-dB gain compression is at +10 dBm minimum. VSWR in/out is 2:1, and powering is +12 to 15 volts. Power is supplied via the output connector by a wideband DC block (bias tee). Standard connectors are type N female. Lightning static protection at the input and reverse polarity protection on the DC supply line are included. Wideband DC block is also available, along with the RX-RX switch (SPDT) and the amplifier bypass switch module (insertion loss 0.15 dB maximum, isolation 50 to 70 dB at 900 MHz, set time 5 ms, through power 10 watt maximum, powering +12 to +15 volts, and VSWR is 1.3:1 typical). The SPDT switch can be used to combine two antennas at the preamp input; the bypass switch module can be used to insert a pad/filter into the signal path.



For additional information contact WI-COMM Electronics, Inc., Box 5174, Massena, New York 13662. Phone: (315)769-8334.

Circle #335 on Reader Service Card.

FAR Breadboards

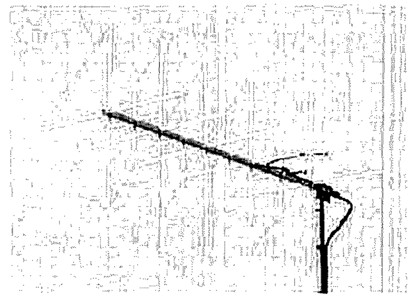
FAR Circuits announces the additon of prototype breadboards to its line of circuit boards. The breadboards come in three different arrangements and can be used for a variety of small circuits and Amateur Radio projects. The cost of the boards is $5 each plus $1.50 per order for shipping and packing. The boards are made of single sided G-10, FR4 glass epoxy material and the copper is solder coated.

For further information write FAR Circuits, 18N640 Field Court, Dundee, Illinois 60118.

Circle #310 on Reader Service Card.

Custom Dual Band Quad Antenna

Custom Antenna Systems now makes a dual-band 2-meter 70-cm quad antenna for the new dual-band radios.



The DB2/70 is a compact, lightweight, high performance beam with five elements for 2 meters and 9 elements for 70 cms. This UHF/VHF dual-band antenna is broad banded and offers a 12.5-dB forward gain on 2 meters and 10.5 dB on 70 centimeters, with front-to-back ratio of 20 dB.

The DB2/70 is 5 feet long and takes a mast size of 1 to 1-3/8 inches. This antenna is end mounted, making it easy to install with only a light rotor. The match system provides low SWR with a 50-ohm feed and a standard PL-259 connector. This beam needs only one feedline, but you may feed both bands separately with a second.

Detailed instructions and precision manufactured components are included for ease of assembly. The boom is constructed of heavy wall 3/4-inch rigid square tubing; the element spreaders are 1/4-inch solid fiber glass rods. The antenna weighs approximately 3-1/2 pounds and will handle a wind load of 90+ mph.

The DB2-70 is available from Custom Antenna Systems. The price is $109.95 plus shipping and handling. Contact Custom Antenna Systems, PO Box 17012, Munds Park, Arizona 86017 or call (602) 286-1236.

Circle #311 on Reader Service Card.

DX-88 HF Vertical Ground Tunable for 80 and 40 meters

Telex® /Hy-Gain® 's new DX-88 design uses the entire antenna on 80 or 40 meters. You can tune 80 or 40 meters to any point on the band without lowering the antenna. You can also adjust the other six bands to any frequency without affecting the tuning of any other band. The DX-88 handles maximum legal power, features unique traps for minimal loss and offers broad-band VSWR of less than 2:1 on six of the eight bands. The self-supporting DX-88 comes with stainless steel hardware and enclosed coils of no. 12 gauge copper wire to reduce loading changes due to weather. With ground radials of 14 feet, the DX-88 requires only a small area for maximum operating efficiency. Optional kits for ground or roof radials, as well as for 160-meter operation are available The DX-88 can also be used as a dedicated SWL antenna and covers 12 bands from 11 to 90 meters. As with all Hy-Gain antennas, the DX-88 comes with a two-year limited warranty.

For detailed information, write to Telex/Hy-Gain, RF Consumer Department, 9600 Aldrich Avenue, South Minneapolis, Minnesota 55420 or call 612-887-5528.

Circle #313 on Reader Service Card.

Toll-free Literature Request Line

ICOM has installed a toll-free literature request hot line for their retail customers. Retail customers can request any ICOM literature by calling (800)999-9877 When you call, a recording will ask you to state your literature request and give your complete address.

FLEA MARKET

RATES Noncommercial ads 10¢ per word; commercial ads 60¢ per word **both payable in advance**. No cash discounts or agency commissions allowed.

HAMFESTS Sponsored by non-profit organizations receive one free Flea Market ad (subject to our editing) on a space available basis only. Repeat insertions of hamfest ads pay the non-commercial rate.

COPY No special layout or arrangements available. Material should be typewritten or clearly printed (**not** all capitals) and must include *full name and address. We reserve the* right to reject unsuitable copy. **Ham Radio** cannot check each advertiser and thus cannot be held responsible for claims made. Liability for correctness of material limited to corrected ad in next available issue.

DEADLINE 15th of second preceding month.

SEND MATERIAL TO: Flea Market, Ham Radio, Greenville, N. H. 03048.

BEGINNER'S RADIO CLEARINGHOUSE. On a space available basis, we are going to offer you, OUR SUBSCRIBER, free of charge, a chance to find a home for your used equipment with a new Ham. Please send us a short description of what you want to sell along with price, name, address and phone number. Please limit your ad to 20 words or less.

FOR SALE: Moving to a smaller home. Wife orders shack cleaning. Completely retiring from Amateur Radio. Selling equipment stored over 20 years. Almost assembly-line new cosmetically. Needs simple rpairs due to long storage. Offering: National NC98, NC125 receivers, NC-5 speaker, FME VHF 152 converter, DB-20 pre-selector. RCA B/C-S/W chassis for console. RCA B/C-F/M ditto, RCA MI series 20 watt pwr amp. Wilcox-Gay CW-3 rack mtd rcvr. All at a fraction of present value. Parts and tubes of all descriptions to meet your needs, receiving *and transmitting, some quite rare. Home brew* receivers for VLF freqs. 500 kHz to 18 kHz with IF's of 370 kHz, 262 kHz, 132 kHz and 85 kHz for the VHF buff, list available. Many parts and tubes FREE to real experimenters and builders especially novices. Tourists and other nearby hams welcome to visit. Call or write George Roberts, 500 West Leeland Hgts Blvd, Lehigh Acres, Florida 33936. Area code 813-369-6724 anytime on low rates. I am close to Ft. Myers.

SIGNAL/ONE CX-11A Transceiver, excellent $2,595.00. Kenwood TSj-930S/AT Transceiver, mint $1295.00. SM-220 $325.00. TSj-711A like-new $750.00. KLM 2M16LBX, new $135.00. Alan, K6GA (714) 964-3912.

FOR SALE: Johnson 1KW antenna tuner, $95.00. Ten-Tec 705-A desk microphone $35.00. B&W 550A 5-position coax antenna switch $20.00. D. Heise, AA6EE, 16832 Whirlwind, Ramona, CA 92065. (619) 789-3674.

STUFF YOUR MAILBOX with Electronic Catalogs, etc. Send $3.00 to Information/HR, 3037 Audrey Drive, Gastonia, NC 28054.

EXTRA CLASS License Test Manuals used once—questions thru Oct. 1991. ARRL $5, Ameco $3 or both $7. Only one each, include SASE in case sold. Nate Williams, W9GXR, 6915 Prairie Drive, Middleton, WI 53562. 100% guarantee to pass if you 100% guarantee to study.

YAGI BUILDERS. Tube traps, tubing, clamps and universal plate for boom to mast or elements. Six band rotary dipole or Yagi. 10thru 40M. SASE for details. 1-800-833-3794, 1-2 PM EST Brown Engineering, Inc, 5501 SW 25th Court, Hollywood, FL 33023.

WANTED FOR MUSEUM: Early microcomputers, SWTPC *6800*, Sphere, Altair MITTS *6800* and other early micros. Also early mirocomputer magazines. David Larsen, KK4WW, Blacksburg Group, PO Box 1, Blacksburg, VA 24063-0001. (703) 231-6478/763-3311.

WANTED: Ham equipment and other property. The Radio Club of Junior High School 22 NYC, Inc. is a nonprofit organization, granted 501(C) (3) status by the IRS, incorporated with the goal of using the theme of Ham Radio to further and enhance the education of young people nationwide. Your property donation or financial support would be greatly appreciated and acknowledged with a receipt for your tax deductible contribution. Meet WB2JKJ and the "22 Crew" at Knoxville, TN Hamfest on June 2. Joe Fairclough will be the featured speaker telling the incredible story of the first *10 years of education thru communication at the core of the* Big Apple. Write us at: PO Box 1052, New York, NY 10002. Round the clock hotline (516) 674-4072.

WANTED: Vibroplex Lightning Bug semi-automatic key (circa 1955). Nate Williams, W9GXR, 6915 Prairie Drive, Middleton, Wisconsin 53562.

BATTERY PACK REBUILDING: Don't pitch it—mail it—for FAST—PROFESSIONAL Rebuilding! Satisfaction guaranteed! ICOM: BP2/BP3 $19.95, BP5 $25.95, BP7/BP8/BP70 $32.95. KENWOOD: PB21 $15.95, PB21H $21.95, PB25/25H/26 $24.95. YAESU: FNB9 $19.95, FNB10 $23..95, FNB4/4A $36.95, TEN-TEC: $24.95 "U-DO-IT INSERTS" ICOM: BP3 $16.95, BP5 $22.95, BP7/8 $27.50, KENWOOD: PB21 $12.95, PB24/25/26 $19.95, TEMPO: S1-15/SERIES $22.95, YAESU: FNB4/4A $32.95, FNB10 $18.95, AZDEN: 300 $19.95, "NEW PACKS" ICOM: BP5 $34.95, BP8S $58.95, YAESU: FNB2 $19.95, FNB10S/FNB12 $44.95, SANTEC: 142/1200 $22.95. TELEPHONE/PAGER/COMPUTER/COMMERCIAL PACKS—Free catalog. Add $3.00 shipping/order. PA +6% VISA/MC add $2.00. CUNARD, RD 6, Box 104, Bedford, PA 15522. (814) 623-7000.

WANTED: 2 meter transceiver base or mobile. Wanted: Interface for TRS-80 computer. WB6VOW, Bob Tibbits, 11023 Marcia Weldon, CA 93283. (619) 378-3146.

VIBROPLEX COLLECTOR'S GUIDE. W1IMQ's new illustrated reference. Key Histories, identification guide, patents, more. 87 pages. Only $14.95 plus $2.00 s/h. Free info. Artifax Books, Box 88-H, Maynard, MA 01754.

TRI-BAND BEAM and tower complete with quiet hilltop custom contemporary QTH situated on 4-1/2 +acres in Francestown, NH. Near ski areas, golf and fine dining. $218,900. Call Norwood Realty (603) 673-8000.

EIMAC 8877/3CX1500A7 new w/warranty $695.00. Eimac Socket/Chimney, for 8877 $125.00. Jennings UCS-300-7.5 KV vacuum capacitor, variable $135.00. RJ1A-26 vacuum relay $75.00. Cardwell 2100PF-3KV variable capacitor $60.00. RF Inductor 25uH-15A $60.00. All—NEW. Alan, K6GA (714) 964-3912.

SLEP SPECIALS: Plate transformers for amplifier builders or replacement gonset P/N271-107 for Models 903, 913 linear amplifiers, 115VAC/1850V at 500MA, size 5-1/4L x 4-1/4W x 4-1/2H, Wt. 18 lbs, new $37.00. HP651B solid state audio oscillator, covers 10HZ thru 10MHz calibrated output, $185.00. Military SG-66/ARM-5 aircraft omni signal generator, mil/version of ARC H-14, ideal for aircraft radio repair $295.00. Military USM-207 frequency counter, late model solid state, eight digit readout 0-500MHz, high stability crystal oven oscillator, size 19W x 5H x 71D, lab quality $185.00. Military TS-510A signal generator 10MHz thru 420MHz, calibrated output, MIL/Version HP608D $195.00, HP616A microwave signal generator 1.8GHz thru 4.2GHz $275.00, have quantity, all lab calibrated/guaranteed, VISA/MC or check, add shipping. Write/phone Bill Slep (704) 524-7519. SLEP ELECTRONICS COMPANY, Highway 441, Otto, NC 28763.

FOR SALE: *Santek ST-200 ET 2m Handheld. 2 battery packs,* DC1 12V converter, headset with PTT control $165.00. Also IC-3SAT, NEW in box $370.00. Daytime phone (603) 878-1441. Will ship UPS. Ask for Marty NB1H.

"SOFTSTART". Protect your valuable tubes and power supply diodes from DANGEROUS inrush surge currents. Can retrofitted into most amplifiers. *Completely assembled Ama*teur net for "SOFTSTART" is $49.95 plus $5.00 shipping in the 48 states. OMEGA ELECTRONICS, 4209 Live Oak Road, Raleigh, NC 27604. (919) 231-7373.

FREE LIST of low cost ham equipment. Write to Jim Braddy, WA4DSO, 3037 Audrey Drive, Gastonia, NC 28054.

WANTED: BUY & SELL. All types of Electron Tubes. Call toll free 1-800-421-9397 or 1-612-429-9397. C & N Electronics, Harold Bramstedt, 6104 Egg Lake Road, Hugo, MN 55038.

HAM SOFTWARE IBM/Compatibles 10 disks $26.95. MC/VISA/Discover. N5ABV EAPCO/H, Bx 14, Keller, TX 76248-0014. (817) 498-4242. 1-800-869-7208.

CHASSIS AND CABINET KITS. SASE. K3IWK, 5220 Harmony Grove Rd, Dover, PA 17315.

RTTY JOURNAL published 10 times per year for those interested in digital communications. Read about RTTY, AMTOR, MSO'S, PACKET, RTTY DX and Contesting. Plus technical articles concerning the digital modes. $12.50 per year (foreign higher). RTTY JOURNAL, 9085 La Casita Ave, Fountain Valley, CA 92708.

CUSTOM MADE EMBROIDERED PATCHES. Your design, any shape and colors. Free ordering information, price list and random sample. HEIN, 79602 Manitou Trail, Glen Arbor, MI 49636-9711.

IBM-PC RTTY/CW. New CompRtty II is the complete RTTY/CW program for IBM-PC's and compatibles. Now with larger buffers, better support for packet units, pictures, much more. Virtually any speed ASCII, BAUDOT, CW. Text entry via built-in screen editor! Adjustable split screen display. Instand mode/speed change. Hardcopy, diskcopy, break-in buffer, select calling, text file transfer, customizable full screen logging, 24 programmable 1000 character messages. Ideal for MARS and traffic handling. Requires 256k PC or AT compatible, serial port, RS-232C TU. $65. Send call letters (including MARS) with order. David A. Rice, KC2HO, 144 N. Putt Corners Rd, New Paltz, NY 12561.

BEAT THE COLD! Melbourne, FL QTH: 4:2:2 (3100 sq/ft) house, 1.6+ acres, 220V wired shack with coax races built in, RHON/25 pad, workshop, pool, sprinkler system, in the country no restrictions, many tall pines and oaks, easy commute to Cape Canaveral, close to fishing/beaches/shopping, SAE for details and photos, $209,000. NODH/4.

AVANTEK ATF10135, $12.00, MMIC's, PC. board, SASE: WA3IAC, 7148 Montague St, Philadelphia, PA 19135.

"HAMLOG" COMPUTER PROGRAM. Full features, 18 modules. Autologs, 7-band WAS/DXCC. Apple, IBM, CP/M, KAYPRO, Tandy, C128 $24.95. HR-KA1AWH, POB 2015, Peabody, MA 01960.

HF PACKET IMPROVEMENT—Digicom 64 revision "A" circuit now available. Board plugs directly into cassette port or remote mount via cable, both connectors included. Power derived from computer. Uses 7910 chip—no alignment required. Switch allows HF or VHF operation. Order Kit #154 for $49.95 or Assembly #k154 for $79.95, both include FREE DISK. Add $3.50 s/h, CA add 6.25% tax. A & A Engineering, 2521 W. LaPalma #K, Anaheim, CA 92801. (714) 952-2114. MC or VISA accepted.

IMRA International Mission Radio Association helps missionaries. Equipment loaned. Weekday net, 14.280 MHz, 1-3 PM Eastern. Nine hundred Amateurs in 40 countries. Rev Thomas Sable, S.J., University of Scranton, Scranton, PA 18510.

HOW TO DESIGN AND BUILD YOUR OWN COMPUTERIZED PRODUCT! New manual makes it easy! Simple step-by-step details from idea to finished product. Includes introduction to design, choosing the proper CPU, comparison between many CPUs & microcontrollers, circuit layout, sample schematics, programming, debugging, hints, tips, and more! $8.00 complete. Satisfaction guaranteed. Details free. TRAVCO, 107 E. Vallette, Suite 1301M, Elmhurst, Illinois 60126.

HAM QTH. Hilltop woodland. Approx. 1500 ft. above sea level. 17.8 acres. Located in NH near Mass. border; perc-tested for single residence, solar exposure, privacy, access road is town-owned, owner-maintained. Phone and power lines within approx. 2500 ft. $79,000, Call or write for details. (603) 547-2053. C.W. Farr, Broker, Rt 1, Box 223, Greenfield, NH 03047.

WANTED: Henry VBC 3000. Dick (716) 386-4092.

INTERESTED IN PUBLIC SERVICE? Join your Local Radio Emergency Associated Communications Team. In Pennsylvania call (717) 938-6943 or write REACT, 1160 Old Trail Rd, Etters, PA 17319.

RUBBER STAMPS: 3 lines $5.00 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

ELECTRON TUBES: Receiving, transmitting, microwave... all types available. Large stock. Next day delivery, most cases. DAILY ELECTRONICS, PO Box 5029, Compton, CA 90224. (213) 774-1255.

RECONDITIONED TEST EQUIPMENT $1.25 for catalog. Walter, 2697 Nickel, San Pablo, CA 94806.

COMING EVENTS

Activities — "Places to go . . ."

SPECIAL REQUEST TO ALL AMATEUR RADIO PUBLICITY COORDINATORS: PLEASE INDICATE IN YOUR ANNOUNCEMENTS WHETHER OR NOT YOUR HAMFEST LOCATION, CLASSES, EXAMS, MEETINGS, FLEA MARKETS, ETC, ARE WHEELCHAIR ACCESSIBLE. THIS INFORMATION WOULD BE GREATLY APPRECIATED BY OUR BROTHER/SISTER HAMS WITH LIMITED PHYSICAL ABILITY.

MAY 5-6: SOUTH CAROLINA: Greenville Hamfest sponsored by the Blue Ridge ARS, American Legion Fairgrounds. Sat 8 to 5. Sun 8 to 3. Admission $4/advance; $5/gate. For tickets or information SASE to Blue Ridge ARS, POB 6751, Greenville, SC 29606.

MAY 5: NEW YORK: The Putnam Emergency Amateur and Radio League will have their PEARLFEST, John F. Kennedy Elementary School, Foggintown Road, Brewster. 9 AM to 4 PM. Rain or shine. Admission $3. For registration please contact Terri Cullum, N2GWF, 40 Mile Hill Road, Highland, NY 12528 or Jim Ward, N2EGS, 22 Trout Place Rd, 4 Mahopac, NY 10541.

MAY 5: MINNESOTA: The Paul Bunyan ARC's annual Hamfest, VFW Club, Bemidji. 8 to 3:30. For information write Carol Johnson, KA0AJD, 1503 Jefferson Ave SW, Bemidji, MN 56601. (218) 751-7920.

MAY 6: MASSACHUSETTS: NOBARC Mayflea, Dalton American Legion, Rt, Dalton. Rain or shine. Starts at Dawn. Admission $1. For information call (413) 458-8452.

MAY 5: NEW YORK: Southern Tier Hamfest sponsored by the Southern Tier ARC, Marvin Park Fairgrounds, Rt 17C—Ex 64, Owego. 8 AM to 4 PM. Admission $4/gate; Tailgate $2/extra; Tables $15; Banquet $15/advance. Talk in on 146.16/76 or 146.52/52. Contact STARC, PO Box 7082, Endicott, NY 13760.

MAY 5: MINNESOTA: The Arrowhead Radio Amateur Club presents Swapfest '90, First United Methodist Church, 230 East Skyline Parkway, Duluth. 10 AM to 3 PM. Admission $4. 4' tables $5. Talk in on 146.34/94. for information contact Duane Flynn, KB0LC, 4907 Peabody Street, Duluth, MN 55801. (218) 525-4580.

MAY 12: MISSOURI: The 15th annual Columbia Hamfest '90 sponsored by the Central Missouri Radio Association, Midway Expo Center 2 miles west of Columbia. Flea Market opens 7:30 AM. Expo Center 9 AM. Tickets 3/$7.50 advance; $3/each at the door. For tickets Ben Smith, K0PCK, 3301 Sinclair, Rt 3, Box 196A, Columbia, MO 65203. Information (314) 443-5168.

MAY 13: OHIO: The Athens County ARA's 11th annual Hamfest, City Recreation Center, Athens. 8 AM to 3 PM. Admission $4. Spouses admitted free. For information write Carl J. Denbow, KA8JXG, 83 Morris Avenue, Athens, Ohio 45701.

MAY 12: WISCONSIN: The Tri-County ARC's annual Hamfest, Jefferson County Fairgrounds, Jefferson. 8 AM to 2 PM. Admission $2.50 advance; $3/door. For information, tickets or tables SASE to TCARC, PO Box 112, Jefferson, WI 53549.

MAY 12: IOWA: The Clinton ARC's Hamfest 90, Iowa National Guard Armory, 1200—13th Avenue North, Clinton. Gates open 8 AM. Tickets $3/advance; $4/gate. Tables $5. Talk in on 145.43 repeater. For information, tickets or tables contact Darryl Petersen, KD0PY, RR 1, Box 84, Bryant, IA 52727. (319) 682-7359.

MAY 19: ARKANSAS: Hamfest 90. Sponsored by the Northwest Arkansas ARC, Community Building, Rodeo Grounds, Springdale. Free admission. For information contact Mike Lorenz, N8FJJ, Rt 2, Box 160A, Prairie Grove, AR 72753. (501) 846-2516 evenings.

MAY 19: COLORADO: Swapfest by the Pikes Peak Radio Amateur Association, Rustic Hills Malls, Colorado Springs. No entry fee. Wheelchair accessible. For information contact Rick, WB7THT (719) 599-7665 or write PO Box 16521, Colorado Springs, CO 80935.

MAY 19: PENNSYLVANIA: Lancaster County Hamfest sponsored by the Ephrata Area Repeater Society, Ephrata Senior High School, 803 Oak Blvd. Genral admission 8 AM. Exams 9 AM. Entrance fee $4. Tailgating $3. Inside tables $6. For infomation and reservations call Tom Youngberg, K3RZF (215) 267-2514 after 6 PM or write EARS, 906 Clearview Ave, Ephrata, PA 17522. All sites handi accessible.

MAY 20: NEW YORK: Long Island Mobile ARC's Hamfest, New York Institute ot Technology, Rt 25A, Old Westbury. 9 AM to 4 PM. Admission $5/gate, no advance. Exhibitors $8. Contact Neil Hartman, WE2V (516) 462-5549 or Mark Nadel, NK2T (516) 796-2366.

MAY 20: NEW JERSEY: Bergen Amateur Radio Assn's Hamfest, Bergen Comm. College, 400 Paramus Rd, Paramus. Contact Jim Joyce, K2ZO, 286 Ridgewood Blvd No., Westwood, NJ 07675. (201) 664-6725.

MAY 20: ILLINOIS: The annual Hamfest sponsored by the Kankakee Area Radio Society, Will County Fairgrounds, Peotone. 8 AM to 2 PM. Indoor flea market and exhibitor tables (limited). Large outdoor flea market. ARRL booth. free parking. Food and drink available. Admission $2.50 advance. $3.00 at the door. Setup kMay 20, 6-8AM. Talkin on 146.34/94. More information from KARS, c/o Frank DalCanton, KA9PWW, RR1, Box 361, Chebanse, IL 60922. Tel (815) 932-6703 after 4 PM CST or (815) 937-2452 before 4 PM CST.

MAY 19: MICHIGAN: Wexaukee ARA Hamfest, Cadillac Middle School, 500 Chestnut St, Cadillac, 8:30 to 2:30 PM. Admission $3. Tables $6. Talk in on 146.38/.98 repeater. Contact John Craddock, KX8Z (a616) 797-5491 or Wexaukee ARA, PO Box 163, Cadillac, MI 49601.

MAY 19-20: WASHINGTON: Yakima ARC, W7AO, celebrating its 60th anniversary, Queen Gym, 5502 West Chestnut Avenue, Yakima. Saturday 9-4, Sunday 9-1. Admission $5 preregistration. $6/door. Features visiting Soviet Amateurs, swap-n-shop, exhibits and more. Talk in on 146.06/66. Contact Yakima ARC, PO Box 9211, Yakima, WA 98909. Mary Wildman, KB7AMF (509) 248-5007.

MAY 20: PENNSYLVANIA: The Warminster AR's 16th annual Hamfest, Middletown Grange Fairgrounds, Penns Park Rd, Wrightstown. Starts 7 AM. Admission $4. Spouse and kids free. For information contact Bill Cusick, W3GJC (215) 441-8048.

MAY 20: WEST VIRGINIA: 12th annual TSRAC Wheeling Hamfest/Computer Fair, Wheeling Park. 8 AM to 3 PM. Admission $2/advance; $3/gate. Free admission 16 years and under. Contact: TSRAC, Box 240, RD 1, Adena, OH 43901. (614) 546-3930.

MAY 26: NORTH CAROLINA: The 16th annual Dur-Ham-Fest sponsored by the Durham FM Association, under the south parking deck of the South Square Mall, Durham. Rain or shine. FCC exams. For information contact Sid Edwards, W4QWM, 1700 High St, Durham, NC 27712.

MAY 27: ILLINOIS: The Chicago ARC will hold its annual Hamfest and Auction, DeVry Institute of Technology, 3300 N. Campbell, Chicago. For details and reservations call (312) 545-3622.

MAY 27: MARYLAND: The Maryland FM Association's annual Memorial Day Hamfest, Howard County Fairgrounds, Rt 144, West Friendship. 8 AM to 3 PM. Donation $4. Tailgating $3. For reservations and information write Melvin Seyle, WA3KZR, 15809 Pointer Ridge Drive, Bowie, MD 20716. (301) 249-6147.

JUNE 2: TENNESSEE: The 24th annual Knoxville Hamfest and Computer Fair sponsored by the Radio Amateur Club of Knoxville, Knoxville Convention Center at the World's Fair Park. 8 AM to 5 PM. Admission $5. Table reservations Frank Ambrister, N4OQJ, PO Box 9605, Knoxville, TN 37940. (615) 933-2539. License exams contact Ray Adams, N4BAQ, 4325 Felty Drive, Knoxville, TN 37918. (615) 687-5410.

JUNE 2: MAINE: Pine State ARC's Hamfest, Hermon Elementary School, Billing Road, Hermon. 8 AM to 2 PM. Admission $2. Contact Roger W. Dole, KA1TKS, RR 2, Box 730, Bangor, Maine. (207) 848-3846.

JUNE 3: PENNSYLVANIA: The 36th annual "Breeze Shooters" Hamfest- NEW LOCATION— Butler County Farm Show Grounds, Rt 68, Butler. 8 AM to 4 PM. Admission $1/gate. Free tailgate Flea Market. Talk in on 147.96/.36. Contact H. Rey Whanger, RD 2, Box 8, Cheswick, PA 15024. (412) 828-9383.

JUNE 3: VIRGINIA: Manassas Hamfest sponsored by the Ole Virginia Hams ARC, Prince William County Fairgrounds. Opens 8 AM. Admission $5. All wheelchair accessible. For information contact Rosemary, KI4VO (703) 361-5255.

JUNE 3: MICHIGAN: 13th annual Swap 'N Shop sponsored by the Chelsea ARC, Chelsea Fairgrounds. Handi parking. Donation $2.50/advance; $3/door. For information SASE to Robert Schantz, 416 Wilkinson St, Chelsea, MI 48118. (313) 475-1795.

OPERATING EVENTS

"Things to do . . ."

MAY 6: The Columbia ARC will operate N4QSL from 1500Z May 5 to 2100Z May 6 from the SC state capital grounds in conjunction with the annual Mayfest Celebration. QSL with SASE to CARC Mayfest, POB 5802, Columbia, SC 29250.

MAY 11: The Escondido Amateur Radio Society will operate a special event station in conjunction with the Annual Avocado festival. Listen for WA6YOO from 0000Z May 11 to 2400Z May 13. For commemorative certificate send QSL and large SASE to EARS, 2435 Our Country Road, Escondido, CA 92025.

MAY 19: The 41st annual ARMED FORCES DAY Communication Test. The Military-to-Amateur crossband operations will be conducted from 19/1300 UTS to 20/0245 UTC.

THROUGHOUT 1990 the Major Armstrong Memorial Amateur Radio Club (MAMARC) will sponsor events commemorating Major Edwin Howard Armstrong's achievements in the field of radio broadcasting. The club is seeking other Amateur operators around the world who are willing to research Major Armstrong's accomplishments and become official MAMARC special events stations. Major Armstrong was a pioneer responsible for the creation of Wideband FM and the inventor of the superhetrodyne receiver. If you are intetested in participating and becoming an official MAMARC special event station contact Barry Groupp, N2HDW, MAMARC, c/o 100th Birthday Committee, PO Box 581, Alpine, NJ 07620. Please SASE.

YOUTH LINK NET. Open to all Hams under age 18. Saturdays at 2000 UTC, 28.425 MHz. For more information contact Net Control, George Manning, WB5NMH, 602 Glendale St, Burkburhett, TX 76354.

FREE 1989-90 Florida two meter repeater directories are currently being distributed by the Hernando County Amateur Radio Assn. of Brooksville, FL. Ask for one at any official Florida Welcome Center or SASE to Repeater Directory, Hernando County ARA, POB 1721, Brooksville, FL 34605-1721.

AMATEUR EXAMS. May 19, July 14, September 8, November 17. St. Mary Medical Center, 3333 No. Seminary Street, Galesburg, IL 61401. 12 Noon to 2 PM. For information contact Larry Heller, KA9PCU, 1436 Brown Avenue, Galesburg, IL 61401. (309) 342-5977.

Monthly Ham Exams. The MIT UHF Repeater Association and the MIT Radio Society offer monthly ham exams, all classes Novice to extra: next-to-last Wednesday of each month, (May 23) 7:30 pm, MIT Room 1-150, 77 Mass Avenue, Cambridge, MA. reservation requested a couple of days in advance, walk-ins welcome; call the shack (617) 253-3776, or Nich Altenbernd (617) 437-0320. Exam fee $4.95. Bring copies of your current license (if any) and Certificates of Completion (if any), two torms of picture ID, and a completed form 610, available from the FCC: (617) 770-4023.

LAUREL ARC monthly (except December) Amateur exam sessions for all license classes. No fee is charged. Preregistration is required. Call (301) 725-1212, Maryland Radio Center, 8576 Laureldale Drive, Laurel, MD 20707.

AMATEUR RADIO CLASSES: For those people interested in obtaining a Novice (basic level) Ham license or upgrading to Tech/General, the Chelsea Civil Defense, in cooperation with QRA Radio Club, will sponsor Amateur Radio Communications classes evenings at Chelsea High School starting MARCH 7, 1989. For more information write Frank Masucci, K1BPN, 136 Grove Street, Chelsea, MA 02150. Please enclose your telephone number.

JUNE 1990 / $2.95

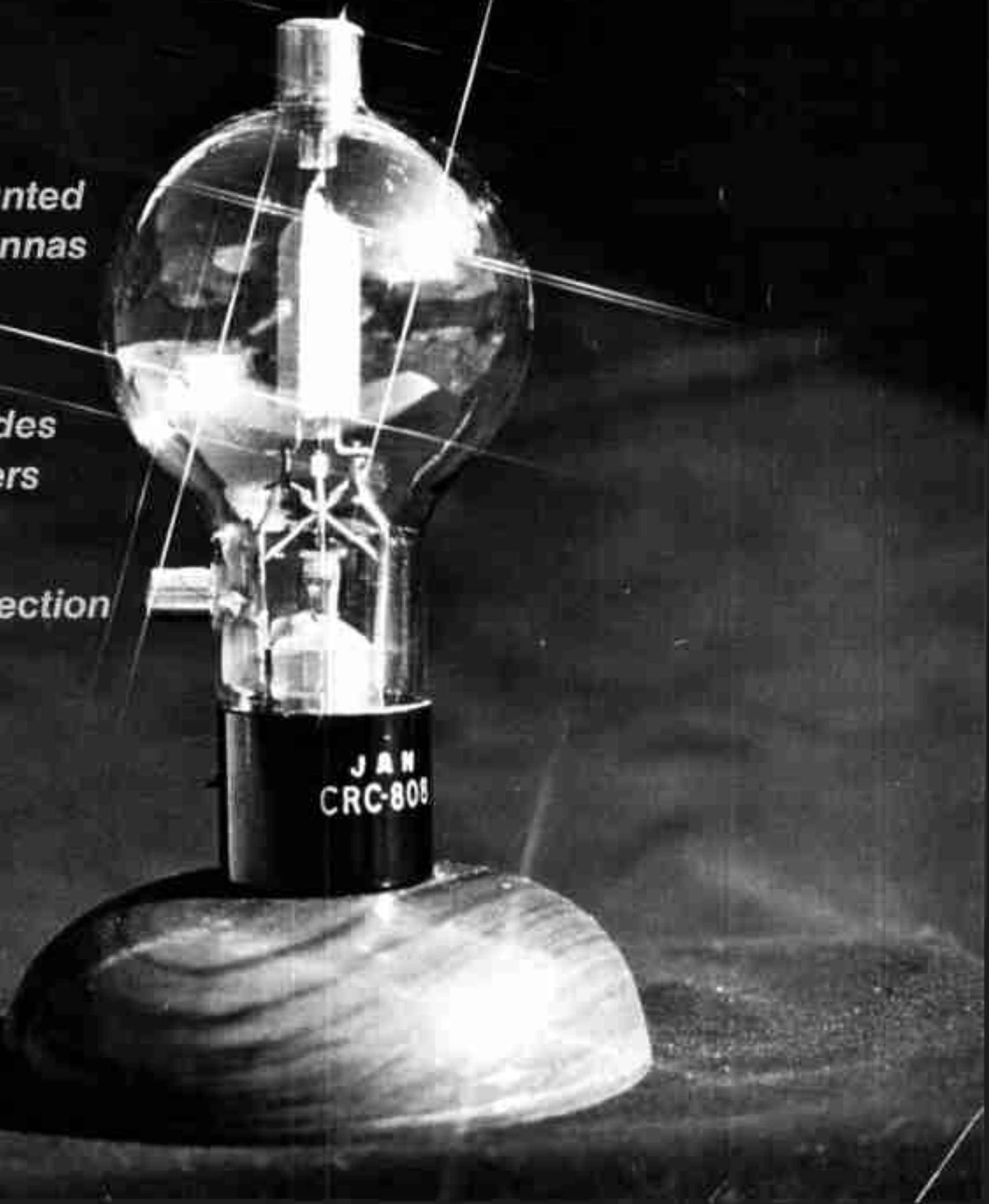
Special this Month:
Three Weekender Projects

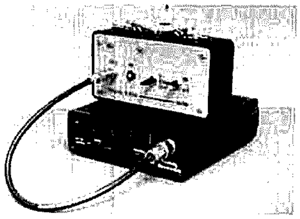
HAM RADIO

Ground-Mounted Vertical Antennas

Varactor Diodes from Rectifiers

The Series Section Transformer





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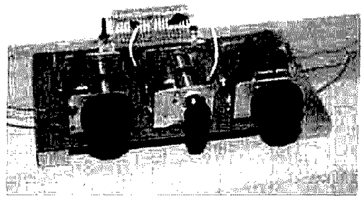

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HAM RADIO

JUNE 1990

Volume 23, Number 6

Cover photo: Michael A. Covington, N4TMI.

FEATURES

DEPARTMENTS

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Microfilm copies are available from Buckmaster Publishing Mineral, Virginia 23117. Cassette tapes of selected articles from **HAM RADIO** are available to the blind and physically handicapped from Recorded Periodicals, 919 Walnut Street, Philadelphia, Pennsylvania 19107.

Second-class postage paid at Greenville, New Hampshire 03048-0498 and at additional mailing offices. *Send change of address to HAM RADIO, Greenville, New Hampshire 03048-0498.*

A New Direction

I have just returned from what turned out to be my most difficult trip ever to the Dayton HamVention™. During the course of the weekend we announced that *Ham Radio* Magazine had been sold to the publishers of CQ Magazine. It was no fun to meet with many hundreds of you and pass on the news that we would no longer be sending our magazine along each month. But it was a very rewarding experience to listen as you told us just how much *Ham Radio* has meant to you.

It's been an extremely productive twenty-two years and I like to think that *Ham Radio* has met most of the goals that the late Jim Fisk, W1HR, and I set for ourselves as we embarked on our new venture back in 1968. As Jim pointed out in his first editorial, there is a very real danger in just publishing state-of-the-art material if you first don't give your readers a clear grasp of the current technology upon which the new ideas are based. We have worked very hard over the years to ensure that a *Ham Radio* reader would be offered this carefully planned balance of information.

This will be your last issue of *Ham Radio*, the last of 268 consecutive regular issues of the finest technical journal ever published for the Amateur. Beginning next month you'll receive *CQ* Magazine, with the addition of a good measure of the very material for which *Ham Radio* was so well known—the most competent technical articles and projects the Amateur will find anywhere.

Why the change? I began to think pretty seriously about this idea a short while ago when Dick Ross, K2MGA, publisher of *CQ* Magazine, asked if we would be interested in being acquired by his organization. By combining the strengths of each publication, *Ham Radio* and *CQ*, the Amateur community stands to gain. Although we both serve the same audience, our two publications have really been complementary rather than competitive as far as the reader was concerned. One was heavily involved with the technical side of Amateur Radio, while the other was directed mostly to the operator.

Both organizations have their strengths. *CQ* has more of the support services, such as typesetting and in-house subscription fulfillment, that help a publisher lower costs. *Ham Radio* has a very strong marketing and direct mail operation that can be used to add to the strength of the whole operation. The **HAM RADIO** Bookstore is unequaled in our industry. It can play an even greater part if it serves the readers of not one but the two strongest independent Amateur magazines, as well as three other *CQ* Communications monthlies.

I would like to thank the many people who have contributed to the success of *Ham Radio* over the last two decades. I hesitate to start naming names because there are so many, and I don't want to leave anyone out. Certainly, we all owe a tremendous debt to Jim Fisk. He set standards for the entire Amateur Radio publishing industry, standards all of us must continue to work very hard to meet.

The Amateur Radio industry has been very good to us. Without advertising support right from the start, there never could have been a *Ham Radio* magazine. Just as important has been the industry's encouragement to continue our task of keeping Amateurs at the forefront of modern telecommunications.

Last, but not least, have been all of you—our readers. You have done your job superbly. Your letters and phone calls have kept us on our toes and have ensured that we never strayed far from the original ideals that we set for ourselves.

What's ahead for the *Ham Radio* reader? A very exciting future. Although I will not be at the helm, I think that I am not out of place to say that you can expect to see the best of what is good in the magazine you already know molded into an even better product. The expanded *CQ* will incorporate the best efforts of two of the most experienced publishing staffs anywhere in the Amateur world. Terry, KA1STC, and Craig, NX1G, will still be here doing their usual excellent job along with a number of other members of the *Ham Radio* team who have contributed so much to our efforts. But now they will have the resources of the *CQ* organization to further their efforts even more. The results should be outstanding.

Skip Tenney, W1NLB

P.S. *CQ* Publisher Dick Ross, K2MGA, said he would love to hear any of your suggestions or comments about all that is happening and how he might best meet your needs in the future.

Backscatter

No New Novices? NO WAY!

Let's assume for the sake of argument that a code-free Amateur license is inevitable, and such a license will be closely related to what the FCC has proposed in PR Docket 90-55. The FCC's version of a Communicator class license would require a written exam based on — but more comprehensive than — the present Technician exam. At the same time the Novice license would be abolished as an entry into Amateur Radio.

I am firmly convinced that such a code-free license would indeed attract some and perhaps many valuable new members to the Amateur fraternity. I am equally convinced that shutting off the much less technically demanding Novice entry into our service would prove at least as damaging as the code-free license could prove beneficial. My conviction is based on a recent personal experience.

My wife, a very bright lady with a Master's degree in Marketing Communications, is taking a basic electronics course. The course is required by her electronics manufacturer employer, who wants his key non-technical employees to understand the language of their field. After helping her cope with series and parallel resistances, inductive reactance and resonant circuits, and seeing how many hours she spends studying to prepare for this class, I have a new and profound appreciation for the difficulty many of us have experienced — even at the Novice level — in becoming licensed Amateurs!

The things she's learning in her electronics course are comparable to what is required for the present Technician license. If, as the FCC proposes, technical expertise of this level will be required for *any* entry into Amateur Radio, the door will literally be slammed in the faces of thousands of interested but not technically oriented would-be Amateurs. These are people who, once they became directly involved through on-the-air and at-the-bench Novice level activities, would eventually pick up much of the technical knowledge required for upgrading in a much more pleasant environment that that offered in textbooks and classrooms.

One of the principal reasons for the FCC's desire to eliminate the Novice as an entry into Amateur Radio is its determination that there's been a good deal of cheating in the Novice licensing process since Novice enhancement has made Novice privileges so much more attractive. Because there was a stipulation when the legislation enabling us to run the Amateur examination program was passed that there should be no charge for administering Novice exams, it doesn't seem likely that VEC organizations could or would be able to assume the financial burden of a formal Novice examination program. However, there surely must be some way to clean up Novice licensing short of abolishing it!

For example, why not require that Novice exams be administered only by accredited Volunteer Examiners, working on their own, but with the results forwarded to Gettysburg through the accrediting VEC? The overall added workload for the VEC would be minimal. As a VE with the DeVry VEC myself, I can't imagine any of my fellow VEs jeopardizing their reputations or the reputation of the DeVry VEC by entering into a conspiracy to give some lazy would-be Novice a free ride! And that's just one possible solution — surely others can come up with equally good or better ways to restore integrity to Novice licensing.

Whatever we do in our efforts to restore vitality to the Amateur service, let's not shut off one established and vital source of new blood in our attempts to open up a new one!

Joe Schroeder, W9JUV

Comments

Freebanders Versus Licensed Amateurs

Dear HR

In March's "Backscatter," I think Bill Orr, W6SAI, brings up a good point in questioning if a no-code VHF license will do that much to increase the ranks of Amateur Radio. But I can't agree completely that the "freebanders" now operating between 26 and 29 MHz are, in his words, "a lot of happy people enjoying the fruits of Amateur Radio the easy way." I agree that they may be happy people on the air communicating via radio, but I disagree that they are enjoying the fruits of Amateur Radio. At least not all of the fruits. Yes, they may be communicating with other freebanders all over the world and so they have 3 MHz to do it in — but what about when the sunspot cycle changes and that portion of the spectrum is dead most of the time?

We hams simply move down to 15 or 20 meters and continue on. The freebander will have only the freebander down the street to talk to. What about the other fruits Amateur Radio has to offer — like packet, SSTV, ATV, RTTY, traffic handling, coordinated emergency communications, repeater operation, and homebrewing your own equipment? If just talking on the air when the skip's in is all you want, then freebanding may be OK, but if you really want to enjoy the different aspects that radio communication has to offer, then an Amateur Radio license is the only way to go.

My main concern with the freebanders is that they're moving up into 10 meters, and causing a lot of QRM in the CW portion of the band. As long as they stay below 28 MHz, I have no quarrel with them, but when they start *encroaching on our* frequencies, I get quite concerned. If they aren't afraid to move into 10 meters illegally, when 10 is dead, I guess there is nothing to keep them from also moving to 15 or 20 or any other ham band.

On second thought, Bill Orr just may be correct. Maybe they *will* enjoy the fruits of Amateur Radio for free unless we, the Amateur community, get busy and ask the ARRL and the FCC to get their heads out of the sand and do something about it!

Bill Harris, K5MIL,
Roanoke, Texas

More on AM phone

Dear HR

Reading the letter by KH6CC and KH6B ("AM alive and well") in the March issue brought back some happy memories I'd like to share. I received my first ticket, W8QXF, in May of 1937 and couldn't wait to get my rig on the air. It was a 6L6 Tritet oscillator for CW on 80 and 40 meters. The Tritet was a crystal-controlled oscillator that could be tuned to either its fundamental or second harmonic frequency and was not too efficient. I sure raised some blisters making that jewel, soldering with a plumber's iron borrowed from my dad and heated over the gas flame in Mother's stove.

My good old Sky Buddy could receive the 160-meter phone band and I would often listen up there and wish that I too could work phone. Money was scarce and there was no way I could buy a modulation transformer. However, an article about Heising Modulation in *The ARRL Handbook* got me thinking. A visit to the friendly radio repairman, not a ham himself, got me a 20-H filter choke from a defunct AM radio. That filter choke, a new crystal for 160, another 6L6, a telephone mike, and miscellaneous small parts from old radios got me on phone. I spent many happy nights on AM phone with my four-tube (including rectifier) rig. I wonder how many remember those transcontinental hookups where we East Coasters would start a net and absorb newcomers across the continent until we had a clear channel and could work distances not otherwise available in the late evenings?

I feel that we lost something by jumping on the SSB bandwagon. I was KH6RD during 1947 and 1948, operating 10 meters only. This was during one of the sunspot cycles and the QRM in the islands was tremendous. There may have been fewer hams then but a large percentage were on 10 meters and all districts would be open at the same time. A ham in Los Angeles and I decided to experiment with the Very Narrow Band FM using reactance modulators. We used slope detection, with sharp crystal filters, and found that we could reduce frequency deviation to such a low figure that the signal sounded like an unmodulated carrier but was perfectly readable when tuned to the middle of the filter slope. What was more important was that we could copy through the heaviest QRM on the band. I think it's a shame that more was not done in VNBFM for ham use. The equipment was simple, with excellent audio, and used little more space than a CW signal. The modern narrow bandpass filter, with a second filter shifted to put its slope in the middle of the passband, or more elaborate limiters and phase lock loop detection should do a superb job with very narrow frequency shift.

We would all benefit by setting aside a portion of the 10-meter band for use of homebrew equipment only on AM, VNBFM, and CW. Make this sub band available to new licensees as well as nostalgic old timers. Limit power input to 100 watts to keep a level playing field. This could put some of the magic back into the hobby for new recruits. Those who choose not to home brew would still have privileges they now have. These modes are simple enough to be homebrewed by almost anyone and, who knows, might start many a kid on an engineering career as it did with me.

Earl Smith, W1BML,
Groton, Connecticut

GROUND-MOUNTED VERTICAL ANTENNAS

A little history and some theory

By W.J. Byron, W7DHD, P.O. Box 2789, Sedona, Arizona 86336 and F.S. Chess, K3BN, 4946 Manor Lane, Ellicott City, Maryland 21043

Vertical antennas date from the beginnings of radio. They were ground mounted with the exception of one highly successful and innovative system, the "Zepp." Its origin seems to have been forgotten, and the Zepp name has come to mean something quite unrelated to the original. In the twenties, when Amateurs were herded into the spectrum above 1.5 MHz ("200 meters and below"), the Zepp proved both practical and useful as a horizontal antenna. The originals, though, were often several thousand feet in length. They hung straight down from the gondola of a Zeppelin (hence the name) — vertically, of course!

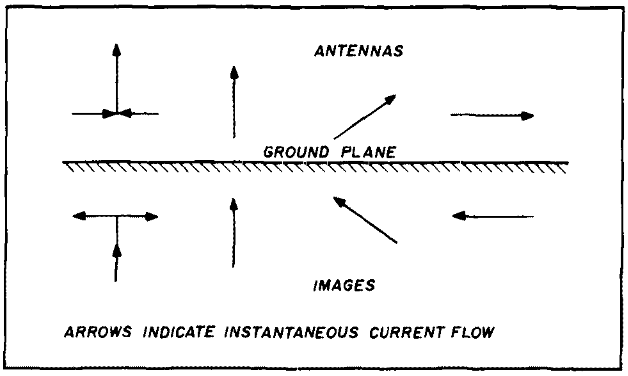
In those very early days, "spark" transmitters were used. They operated on enormous wavelengths (sometimes kilometers), and a number of them were run at many kilowatts with surprisingly high efficiencies — 80 percent was rather common. POZ, at Nauen, Germany operated at 150-kW output, 85-percent efficiency at the transmitter. Because of the extreme shortness of the antenna and the long wavelengths, antenna efficiency was probably less than 5 percent at most, resulting in an antenna current of over 1,000 amperes! There were also "arc" transmitters, and two different types of mechanically-generated radio frequency transmitters, but these appeared a decade or more later.[1]

Why vertical?

Vertical antennas were (and are) the only way to launch a low frequency radio field from a location on or near the ground. The techniques, practices, and experiences of earlier antenna pioneers are still germane today — especially when the subject is verticals. There's not much that's really new after all this time. Our instruments and some techniques have simply been upgraded.

There's a notable, fundamental difference between the behavior of verticals and horizontal antennas operated near ground. The "images" shown in **Figure 1** are in phase for the vertical, but they are opposing in the horizontal case.

FIGURE 1



Current filaments and their images over "perfect" ground.

Thus, while the vertical can be operated right on the ground, the horizontal can't because the image tends to cancel the antenna current. It would cancel completely if there were such a thing as "perfect earth." The perfect earth concept is used extensively for preliminary designs. Ground conditions, when known, aid in final designs.

Because the first transmitting antenna was made by Marconi (apologies to Tesla, Popov, and probably others who could have been first) the base-fed short vertical antenna is called (you guessed it) a "Marconi." It's defined as follows:

A Marconi is a current-fed antenna (usually vertical) whose overall length is a quarter-wave or less, and is loaded by various means so that it exhibits series (90 degree) resonance.

Those "various means" may include series inductors, capacitive top hats, or mixtures of both. Most of the early systems used a combination approach. There were two main reasons for this. The first was that erecting a quarter-wave high antenna for a wavelength of 5,000 meters was impossible then, and would be nearly so today. The second involves the "logarithmic decrement" of the antenna, about which there will be more later. Because it was related to the bandpass — those were *very* broad signals — it was of great importance to control (always to decrease) the value

of the logarithmic decrement, hence the bandpass. In practical terms, this meant that inductance had to be added in order to decrease the log decrement, which was subject to government regulations. In 1919, the United States Department of Commerce limited the log decrement to 0.2 maximum.

Ever narrower bandpass

All spark transmitters functioned by the periodic discharging of a capacitor bank through a spark gap. The spark gap was either in series with an antenna which was the *only* frequency-determining element in the system (Marconi's early method), or coupled to a secondary circuit which contained the antenna and various resonating components — usually inductors. When a charge was delivered to either of these two basic types, the circuit plus antenna (or the antenna itself) would "ring," losing some of its energy with each cycle because of radiation and circuit losses. The decrease in amplitude was a constant fraction of the preceding cycle amplitude. That constant was the logarithmic decrement, and is related to what we now call the "Q" of an antenna — except that it is inverse. The larger the Q, the smaller the log decrement. The main factor controlling the log decrement was the amount of inductance in the system (as with Q, which is X_L/R), thus most of the early verticals were base loaded even though they had very large top hats. It was the only practical way to control the log decrement; the only place you could introduce inductance conveniently was at the base. The large (even by today's standards) top-loading capacitances were the smaller part of the total loading when the wavelength was several kilometers. Some of those early top-loading schemes resulted in 0.05-μF capacitance. The pioneers taught us something very important — how to construct base-loading inductors with an intrinsic Q of 5,000 or more. Such numbers are possible when the frequency is below, say, 100 kHz, where Litzendraht ("Litz") wire is practical. Litz wire loses its effectiveness above a few hundred kHz. Modern OMEGA antenna systems use Litz wire-loading coils; they are also combination base and top loaded.

An about-face

Nowadays we try to increase the bandwidth of a vertical, just to avoid having to retune when a rather large frequency shift is made. The signal bandwidth is *controlled by the modern design of both the transmitter and receiver.* It's a challenge to have a wide antenna bandwidth while maintaining high efficiency. It's especially difficult when the antenna is physically short, as is usually the case with vertical antennas for 80 and 160 meters. This is particularly true when they are placed over good ground systems.

We mentioned before that the antenna current for POZ was over 1,000 A (references exist which mention 1,200 A). Today, using more or less typical transmission line impedances of 50 or 75 ohms, a 1,000-A line current would represent 50 to 75 MW. But we know that the power was "only" 150 kW. Why the large current? Obviously, there were very low impedances involved. A characteristic of LF and VLF Marconi antennas is their low feedpoint impedances, given little ground loss. Their actual radiation resistances were extremely low, usually a fraction of an ohm. A resistance of 50 milliohms was rather common. That's because they were all physically very short.

How do we calculate radiation resistance?

The radiation resistance of a half-wave doublet constructed of "infinitely thin" wire (a necessity for the original derivation) is approximately 73.2 ohms.* That value has been known since the 1880s. Therefore, the radiation resistance of half a very thin doublet would be 36.6 ohms. This is the generally accepted value for a full length quarter-wave vertical antenna, unachievable though it is. We deal with values less than that because the length necessary for resonance is also a function of the length-to-diameter ratio of the element. The outcome is resonant lengths that are shorter than the equivalent "free space" lengths. The net result is an antenna that will always be of lower resistance than that achieved for the thin wire. **Figure 2** shows the manner in which it changes.[2]

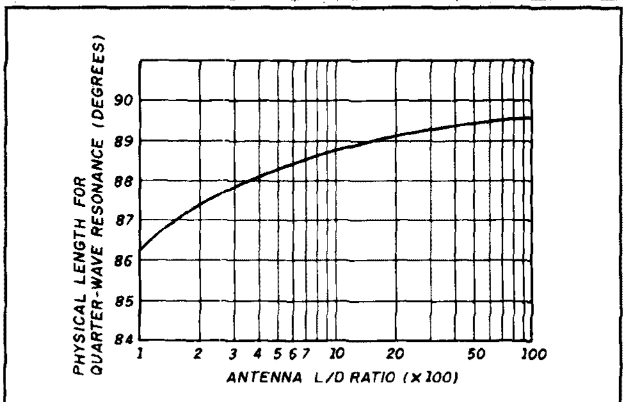
It has been difficult to determine the expected radiation resistances of variously loaded short verticals, although some complicated calculations do exist. Occasionally, a curve is published titled "Radiation Resistance of a Vertical Antenna versus Height," with no indication as to the form of antenna (or what length-to-diameter ratio) it relates to. It's usually presented on linear graph paper, though the function is a steep transcendental one which makes it hard to interpret at both ends of the curve. Typically it's been calculated for what we'll call the "base-loaded" case, and is of no use whatsoever for any other type of loading (like top, center, or combination).

Simple but workable derivations

Figures 3 through 6 illustrate methods for estimating the radiation resistances of various antennas with different forms of loading. Two of them were first presented in *Ham Radio* in 1983.[3] The derivation starts with one-half of the theoretically derived resistance of a free space half-wave dipole (36.6 ohms). If you assign 1 A as the value of the base current and assume that the current distributes itself sinusoidally, then the area of the profile will be 1 ampere-radian.

The rationale for the derivations is that by allowing for no

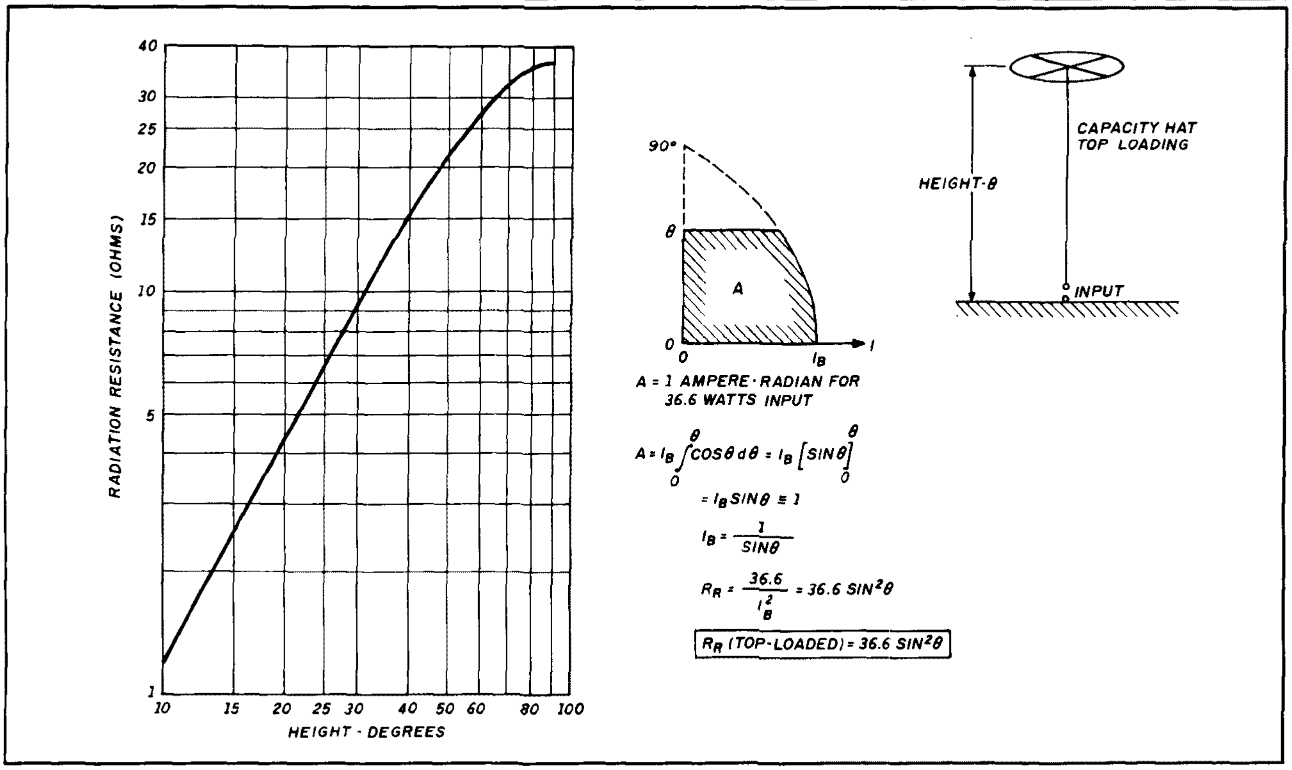
FIGURE 2



Graph of free-space length of quarter-wave resonance versus the ratio of the length-to-diameter of an element.

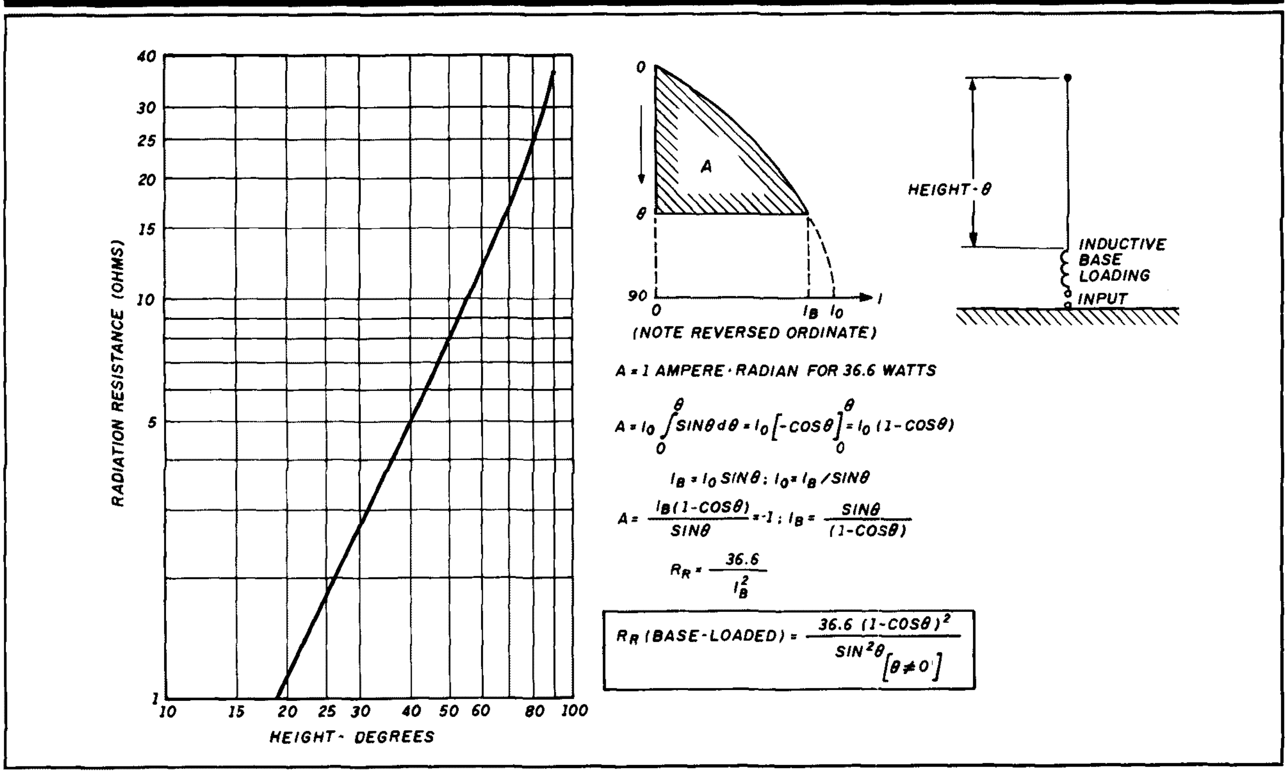
*This information can be found in several sources. We used one of Schelkunoff's antenna books.

FIGURE 3



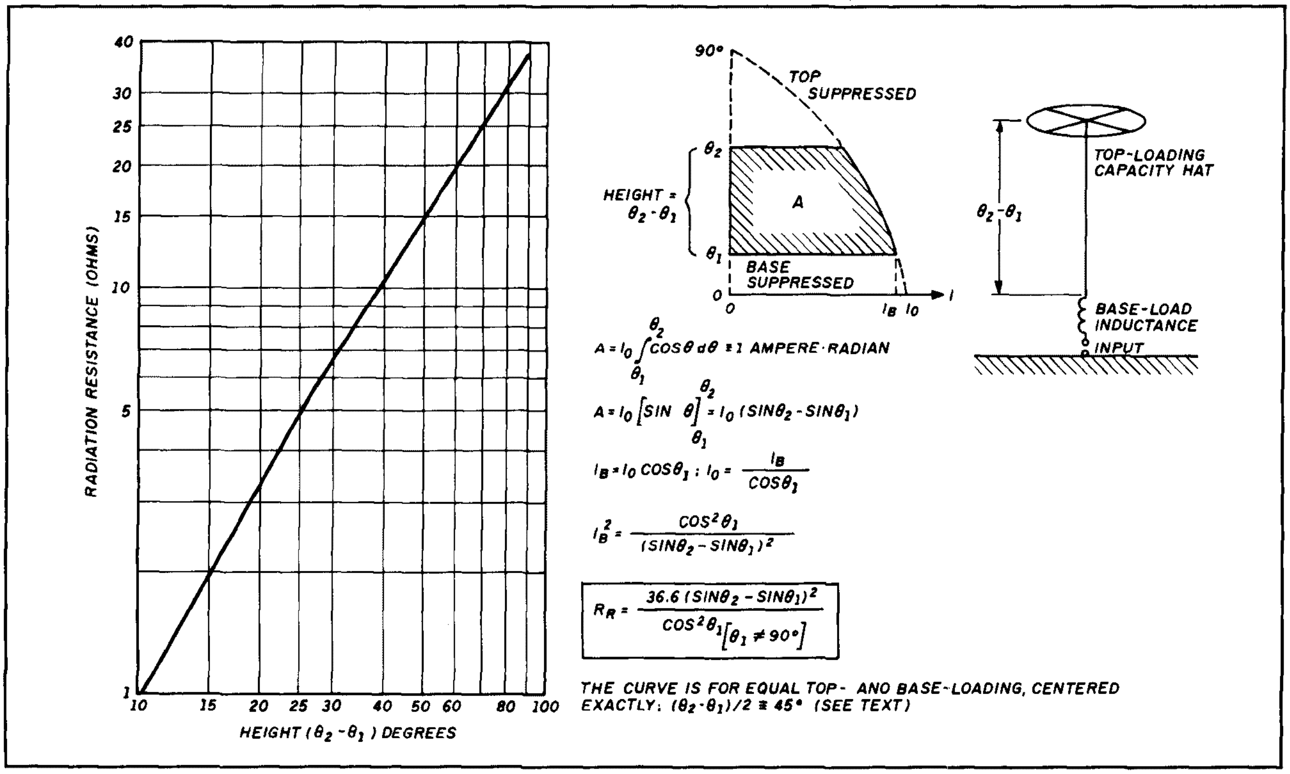
The curve and its derivation for the top-loaded Marconi vertical. The curve is computed from the expression in the rectangular box.

FIGURE 4



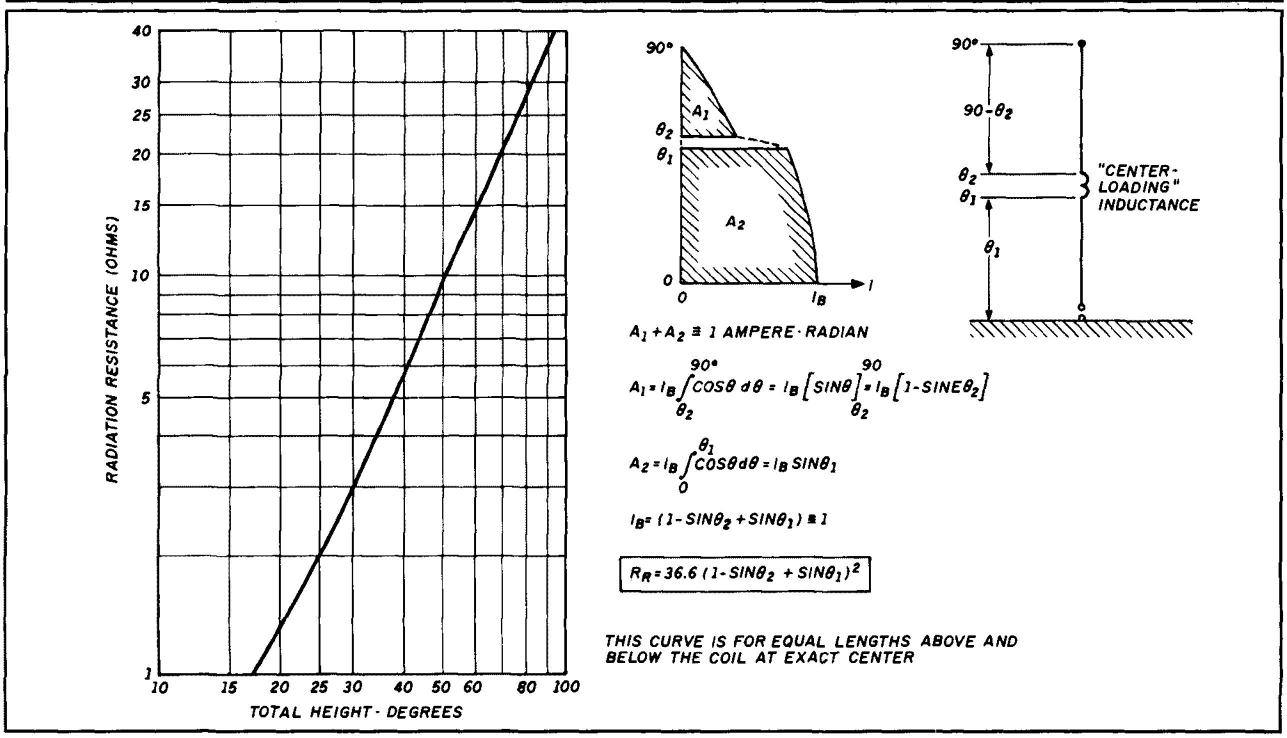
The curve and its derivation for the base-loaded Marconi vertical.

FIGURE 5



The curve and its derivation for a combination base and top-loaded Marconi vertical. This curve is for a combination of equal suppression at the top and bottom. This case and that of the "center-loaded" will produce a family of curves.

FIGURE 6



Curve and its derivation for the center-loaded case. The curve is for a coil in the exact electrical center. Technically this is a "segmented" antenna.

losses other than radiation, any configuration of loading will produce one or more profiles, the sum of which must total 1 ampere-radian for any single antenna. A trigonometric-algebraic expression evolves for the base current in each case. This expression is squared and divided into 36.6 watts, the assumed power input for 1 A in the reference antenna. The quotient is the radiation resistance as a function of the exposed element lengths expressed as angles. These derivations provide very good estimates of the radiation resistances, and can be calculated by anyone who has a handheld scientific calculator.

There are four derivations. The last one, shown in **Figure 6**, presents calculations for the "center-loaded" case. It's of limited use, for two reasons. First, few center-loaded verticals exist in which the top section is the same diameter as that of the bottom — a requirement for the derivation. It's usually a thin "whip." Second, the rules governing the way the derivations were perfomed may not fit the case as well as they do in the first three. It's difficult to determine the amount of "standing wave" that exists on the loading coil, if any at all.* We believe there is some standing wave on the loading coil, and intend to make some measurements to confirm it.

However, the existence of this wave may not be important, because the whip actually contributes almost nothing to the transmitted field. Almost all of the signal comes from the section below the coil. Thus you may estimate the radiation resistance of the antenna by using the expression for the top-loaded vertical with the length set equal to that of the bottom section plus the coil. Those who build a center-loaded vertical would be well advised to use a top hat and *to place the coil — which would be considerably reduced* in value because of the increased top capacitance — directly under the top hat, or to eliminate the coil entirely and resort to top loading alone. It's interesting to note that the coil under the top hat configuration isn't new; it was covered by a United States patent in 1909!

The combination top and base-loaded case (**Figure 5**) was solved for equal loading at the top and bottom; the radiating portion of the antenna is in the exact electrical center. A simple BASIC program for the computation appears as **Figure 7**. Both this concept and that of the center-loaded antenna will produce a family of curves. You can modify the program to generate the rest of the family.

Probably the most important characteristic of the top-loaded versus base-loaded vertical is that for heights up to about 30 or 40 degrees, the radiation resistance of the top-loaded vertical is nearly four times larger than that of the base-loaded system. This means that the top-loaded antenna would be four times as efficient as the base-loaded antenna erected over identical ground systems. We don't recommend using base loading in just any situation, except as a *tuning network or part of a mostly top-loaded combination*. These derivations also revealed that all other combinations of top and base loading result in radiation resistances between those two curves.

In 1977 *Jerry Sevick, W2FMI, published what may be the* best article in recent literature on short ground radials for short verticals.[4] Jerry tested many combinations of short (Marconi) verticals over several radial systems. One of his figures, shown here as **Figure 8**, plots measured values of

FIGURE 7

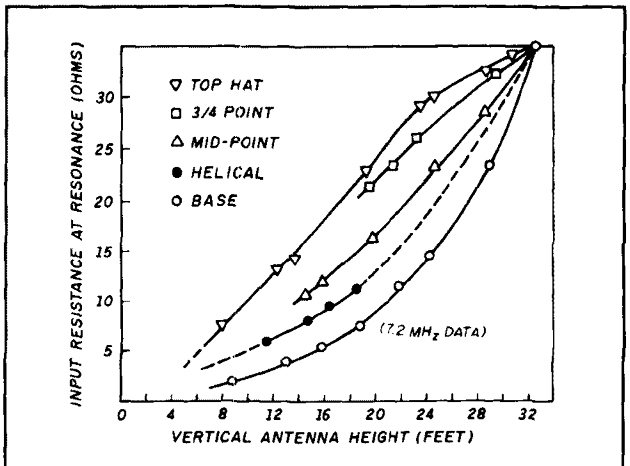
```
10 A = 45:B=45:
20 TH1 = 3.14159*A/180
30 TH2 = 3.14159*B/180
40 M = SIN(TH1):N = SIN(TH2)
50 RR = 36.6*(N-M)^2/(COS(TH1))^2
60 LPRINT USING "###.##     ###.##";B-A,RR
70 A = A-5:B = B+5
80 IF (B-A)=<90 THEN GOTO 20 ELSE END



 0.00     0.00
10.00     0.95
20.00     3.29
30.00     6.54
40.00    10.42
50.00    14.81
60.00    19.61
70.00    24.83
80.00    30.48
90.00    36.60
```

Short BASIC program to calculate the curve in *Figure* 5. The column lists first, the total length (B-A), and the second, its radiation resistance (R). Notice that at 90°, both the top and base loadings disappear, and the resultant is the resistance of a quarter-wave antenna.

FIGURE 8

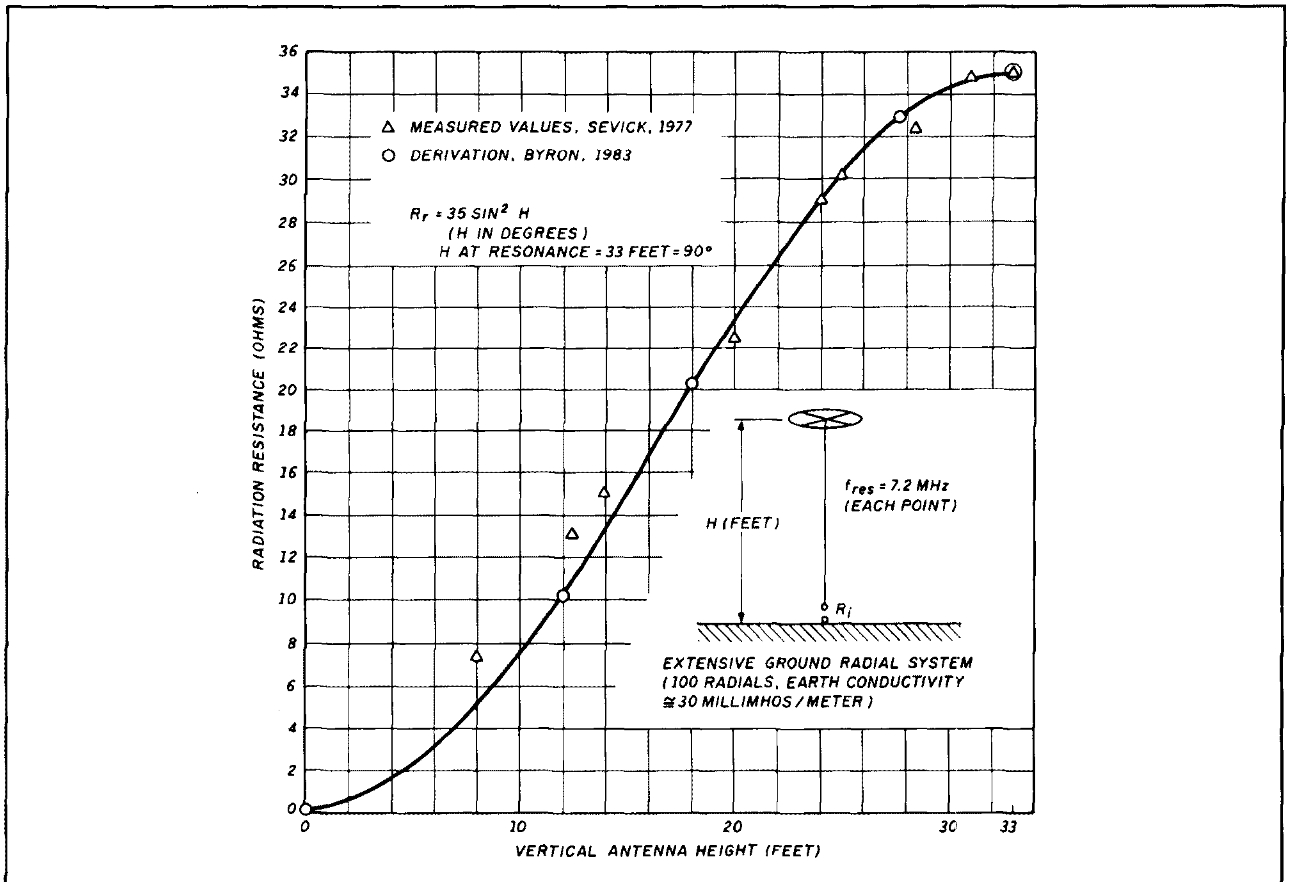


Measured values from Jerry Sevick's (W2FMI) antennas erected over a nearly lossless ground system for several representative types from his 1973 article, "Short Radial Systems for Short Verticals," in *QST*. Used with permission of the author.

input impedance (resistance in these cases) over a very good ground system of 100 quarter-wave terminated radials. The earth conductivity was higher than 30 millimhos per meter; it doesn't get much better than that! We're sure it was a result of Jerry's well-fertilized back lawn. The curve for the top-loaded antenna (the top curve) gives a value very close to its radiation resistance because there's virtually no parasitic ground loss and no coil resistance. The other curves

*Work by Robert Lewis, W2EBS, and the late Edmund Laport corroborates this finding

FIGURE 9



A plot of the measured values from *Figure 8* (the top curve) curve-fit by the derivation from *Figure 3*. As there was little ground-loss, and because there was no inductance, the points for the top-loaded system represent radiation resistances.

reflect the effects of the various loading coils. Even the construction of a coil of large wire (like no. 10) will contribute several ohms of parasitic resistance, depending on the frequency and inductance. This *cannot* be ignored. The RF resistance of a 13.8-μH coil constructed of no. 16 silver-plated copper wire measured 1.6 ohms at a frequency of 7.2 MHz.

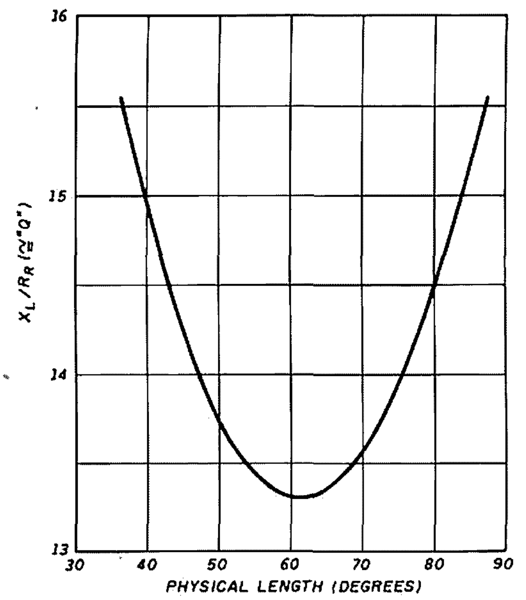
Proof of performance

Figure 9 shows the top-loaded antenna data from **Figure 8** as "curve fit" by the expression for top-loaded verticals from **Figure 3**. We did this to demonstrate that the coefficient (36.6) is just that — a coefficient set as a consequence of this particular derivation method. If the original dipole radiation resistance calculations had yielded, say, 32 ohms, then it would have appeared in the derivations. Jerry's full length vertical measured 35 ohms. This becomes the new coefficient and illustrates how you'd use these expressions. Notice how well the experimental data (from 1977-78) fit the curve (essentially from 1982) as published in 1990, even though a refinement to accommodate the decreasing L/D ratio for the points below about 14 feet in height wasn't performed. Given the absence of measured values for a full length vertical (the usual circumstance), a good starting coefficient for verticals of, say, 1-1/2 or 2 inches diameter would be 35. The arbitrary assignment of 35.0 wouldn't produce an error of more than 3 or 4 percent in almost any case, and probably less. Simply substitute 35 for 36.6 in all the derivations.

A surprise

We've all been told over the years that a shortened antenna results in a narrower bandpass than that of a full-sized vertical. This is quite obviously true for a base-loaded vertical, but might not be true for the top-loaded system. In 1989, Frank Chess made some calculations for top-loaded systems and calculated the inductance of the vertical section. He computed both the inductive reactance (X_L) and the radiation resistance (R_R) as the element was shortened. He assumed it to be over zero-loss ground. Consequently, as the element is shortened, resonance is restored by increasing the size (capacitance) of the top hat. The results are very interesting. As a matter of fact, they're startling. The Q of the vertical *decreases* as it's shortened down to a physical length of a little over 60 degrees, and then it increases (see **Figure 10**). We leave this as conjecture; we haven't performed any experiments for confirmation. However, if this is true, a totally top-loaded vertical over a good ground system resonant at a 62-degree electrical height may have a wider intrinsic bandpass than that of a full-sized quarter-wave vertical for a reasonable length-to-diameter ratio of, for instance, 200 or so.

FIGURE 10



The curve for the "Q" of a top-loaded antenna as calculated by one of the authors (Chess, K3BN) in 1989. As a full-length vertical is shortened, it is assumed to be brought to resonance by a top-hat. The inductive reactance (X_L) of the vertical section is calculated along with its radiation resistance (R_r) in each case.

With few exceptions, the behavior of most antenna systems is influenced by the ground beneath it. hr

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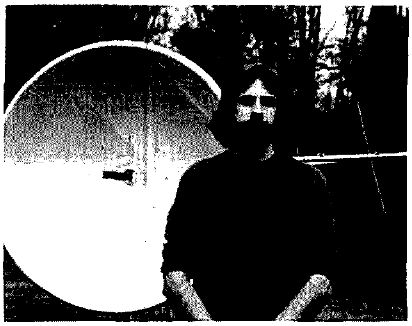
2. Jonathon Zenneck, *Wireless Telegraphy*, translated by A.E. Seelig, McGraw-Hill, 1915. An original derivation was done by Heinrich Hertz in 1888.
3. W.J. Byron, W7DHD, "Short Verticals for the Low Bands," *Ham Radio*, May 1983. See Figures 5 and 6.
4. Jerry Sevick, W2FMI, "Short Ground Radial Systems for Short Verticals," *QST*, April 1978.

Bob Atkins, KA1GT

MICROWAVE BASICS



This isn't the column I had originally planned for this month. Instead I've written in response to a number of letters from Amateurs who want some very basic information on microwaves. Because it's easy for me to forget that not everyone knows about the microwave bands, I'll devote column space this month to an introduction to microwaves. My own personal preference is towards weak signal work, so this may be reflected in my descriptions of microwave activity. However, other modes (like ATV, FM repeaters, and packet) can be found on the bands.

What are "microwaves?" As a working definition, consider microwaves to be electromagnetic waves with a frequency greater than 1000 MHz or a wavelength less than 30 cm. That defines the lower frequency limit, but how high in frequency do microwaves go? The answer to this question is somewhat nebulous; an upper limit would be somewhere around 300 GHz (300,000 MHz). Within this region there's a subdivision referred to as millimeter waves. It's the area between 30 and 300 GHz, where the wavelength is less than 1 cm. United States Amateurs have 11 bands allocated above 1 GHz, with a combined bandwidth of 23 GHz (23,000 MHz), so you can see that there's plenty of room! I'll define and describe these Amateur bands later. At this point, you just need to know that they exist and have an idea of their size.

Before dealing with any of the technical aspects of microwaves, it may be helpful to try to answer a question which has been posed to me quite frequently. What is the attraction of working on the microwave bands? This is a difficult question and probably has as many answers as there are active microwave Amateurs. The microwave bands provide the challenge of exploration. Many of the bands are sparsely populated; indeed some of them have never been worked at all! There's a real opportunity to be "first" on a band, make the first contact over a given path, discover a new mode of propagation, or make a real technological contribution to state-of-the-art operation. Even if you don't do any of these things, there's a sense of personal satisfaction and achievement in being one of the pioneers on the newly developing bands. The low population of the bands lends itself to other characteristics. For instance, there's no QRM, and in my opinion the overall operating standards are higher. Listening to some of the poor conduct on the HF bands is a distressing experience. I've never heard any discourtesy, deliberate interference, or other undesirable conduct on the microwave bands. The low population also encourages a sense of community and cooperation between those Amateurs who are on the microwave bands. It's in everyone's interest to help develop activity — even if it's just so there's someone to talk to on the air! In a way, microwave operation is a throwback to the very early days of ham radio — not in technology, but in spirit.

Now let's move on to some of the technical aspects of microwave operation and see how they differ from those found on the lower bands. Perhaps the most obvious difference is the shorter DX range. While many paths in excess of 1000 km have been worked, such DX is rare. I'll cover this in more detail later, but for now it's enough to know that range will be a few hundred kilometers. Such DX work uses antennas with high gain (>20 dBi), and hence a very narrow beamwidth. Consequently, you have to know where the DX is in order to point the antenna with sufficient accuracy to work it! As a result, a lot of microwave work is based on prearranged schedules, or relies on liaison at lower frequencies (often 144 MHz). Much of the activity on the microwave bands is concentrated during contests, local activity nights, or local nets. For example, the Pack Rats group in the Philadelphia area has a 1296-MHz net every Monday night at 9:30 p.m. on 1296.1 MHz.

The importance of a good QTH site is paramount on the microwave bands (especially at the higher frequencies); thus there's a lot of portable operation from mountaintops. As typical microwave TX power is low (maybe 1 watt), and antennas are small (a 3 or 4-foot parabolic dish on 10 GHz is about the practical limit), portable operation is quite convenient. Photos A and B show examples of portable operation on 3456 and 5670 MHz by members of the North Texas Microwave Society.

For those not interested in working DX, the microwave bands offer one other unique feature — bandwidth, and lots of it. For example, the 10-GHz band is 500 MHz wide and ideally suited for high speed digital data links which can rapidly use up bandwidth. Because microwave antennas are highly directional, and DX propagation is difficult, a number of high speed data links can use the same frequencies, even when they are close to each other geographically. Similar factors apply to other wide bandwidth modes. Who will be the first Amateur to transmit high speed digital HDTV signals on the air? One mode uniquely permitted on the microwave bands is pulse modulation. Though this mode is prohibited on all bands below 1000 MHz, it's allowed on all the microwave bands except for 1240 to 1300 MHz and 10 to 10.5 GHz

As I mentioned earlier, there are 11 Amateur microwave allocations. I'll take a look at each one and try to describe their characteristics. Some of these bands have full or partial ARRL-recommended band plans. In general, narrowband weak signal work takes place on frequencies related to harmonics of 1152 MHz (for historical reasons involving frequency multiplying).

- 1240 to 1300 MHz, or the 23-cm band. This is the most populated

PHOTO A

N5JJZ/5 demonstrates portable operation on 5760 MHz.

PHOTO B

North Texas Microwave Society Expedition on 3456 MHz. Elevation is controlled by jacking up the rear of the van! WA5TNY is seen adjusting the dish.

microwave band, and the one most like the VHF/UHF bands in its operational characteristics. Novice operation is permitted between 1270 and 1295 MHz. Conventional vacuum tubes (2C39) can be used to generate power, and power levels of 100 watts aren't uncommon. Antennas are generally of the Yagi type (or loop Yagi where 1-wavelength loops replace the usual 1/2-wavelength elements like a multi-element quad). Antennas are quite small; a 25-element Yagi is a little over 6 feet long. Fixed station operation is common, and a well-equipped station can expect a DX range of several hundred kilometers under flat conditions. The California/Hawaii path has been worked on this band (3977 km) using modest power and antennas (N6CA's 100-watt, 44-element loop Yagi to KH6HME's 25-watt, 4 by 25-element Yagis). Under good conditions, distances of up to 100 km can be worked with power levels of less than 1 watt. The narrowband calling frequency is 1296.1 MHz. The ARRL band plan has allocations for repeaters, satellite uplinks, ATV, and digital communications, and there's activity on all these modes. Much commercially built equipment is available for this band — from complete multi-mode transceivers costing well over $1000, to low power transverter kits in the $140 price range.

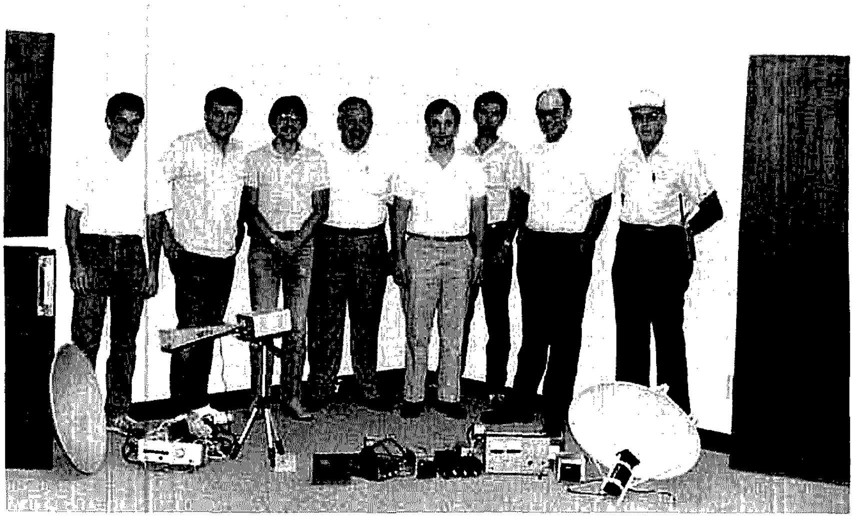
- 2300 to 2310 MHz and 2390 to 2450 MHz, or the 13-cm band. Activity here is much lower than on 23 cm, though it's growing and quite well established in some areas (like North Texas and Philadelphia). Quite a lot of operation takes place from fixed stations. Power can still be generated with cheap vacuum tubes (2C39s, $10 surplus) at levels of 20 to 30 watts. Antennas are usually of the Yagi type, but some parabolic dishes are in use. Range is less than at 23 cm, but considerable distances can be worked with low power under good conditions (W8YIO's 30-mW, 8-foot dish to WA8TXT's 200-μW, 4-foot dish. There were strong signals at 135 km; the path could have been worked using 2-foot dishes). The narrowband calling frequency is 2304.1 MHz. Commercially built equipment is available. The mode S satellite downlink is in this band (2401 MHz).
- 3300 to 3500 MHz, or the 9-cm band. Activity is very low, but growing slowly. Conventional vacuum tubes don't work well at these frequencies and power must be generated using solid-state devices or exotic vacuum tubes, like traveling wave tubes or klystrons. Parabolic dishes are the most common antennas. Much of the equipment is homebrew, though there are now some commercial transverter kits available. Some TVRO (satellite TV) equipment can be modified for use in this band, providing excellent performance at low cost. The narrowband calling frequency is 3456.1 MHz.
- 5650 to 5925 MHz, or the 6-cm band. Activity here is very low. There are a few stations on the band with homebrew systems (often containing surplus commercial parts). Currently, there's no readily available commercial equipment for this band. The narrowband calling frequency is 5670.1 MHz.
- 10 to 10.5 GHz, or the 3-cm band. Second only to the 23-cm band, this microwave band has one of the highest levels of activity. This band is popular because it's easy to get on using wideband equipment based on Gunn oscillators. A Gunn oscillator is a resonant cavity containing a Gunn diode. When a DC voltage (about 9 volts) is applied across the diode, oscillation occurs at the resonant frequency of the cavity. Power output is in the 10 to 100-mW range. Complete commercial wideband systems (Known as "Gunnplexers"™) are available for a few hundred dollars. A homebrew system can be built for a few tens of dollars and not much technical microwave knowledge is required. Antennas are usually parabolic dishes 1 to 3 feet in diameter. Any line-of-sight path can be worked using such equipment, but almost all obstructed paths require the presence of enhanced propagation modes (ducting) which occur infrequently. The world record of 1000+ km was made using such a wideband system. In addition to wideband systems, there are an increasing number of narrowband systems coming on the air which use conventional transverter techniques. This is in large part due to the availability of commercial (SSB electronics) transverter components (local oscillators, transmit and receive mixers). Kits are available for a few hundred dollars; built and tested units cost about double. Power output is in the 100 to 200-mW range, and this is enough to work many obstructed (non line-of-sight) paths of several hundred kilometers on a regular basis. Such paths would be difficult, if not impossible, to work using wideband equipment. Most of the operation is still done from portable stations on hilltops, but a number of stations are developing fixed station capabilities. Several stations have employed ATV and digital modes on this band. The

ARRL has a 10-GHz contest which takes place every year over two weekends in August and September. Last year the leading station (W6HHC) made 78 contacts with 17 different stations and a best DX of 266 km. Thirty-one stations made contacts in excess of 100 km. Oscar 9 carried a beacon on this band (10.470 GHz). The narrowband calling frequency is 10.361 GHz.

- **24 to 24.25 GHz,** or the 1.5-cm band. There isn't much regular operation on this band. The majority of activity involves Gunn oscillator-based systems, though some narrowband transverters have been built. This is the lowest frequency band at which attenuation by the atmosphere (oxygen and especially water vapor) is a significant factor in propagation, amounting to about 0.2 dB/km.
- **47 to 47.2 GHz.** This band has no alternative wavelength name. It's the lowest frequency Amateur millimeter wave allocation. Atmospheric attenuation is significant at about 0.4 dB/km. There is Amateur activity, but it's confined to one or two stations.
- **75.5 to 81 GHz.** This is the highest frequency band on which I am aware of Amateur activity. Atmospheric absorption amounts to about 0.4 dB/km.
- **119.98 to 120.02 GHz.** There's no known Amateur operation at this level. Atmospheric absorption is around 2 dB/km.
- **142 to 149 GHz.** No known Amateur operation — there's atmospheric absorption of around 2 dB/km.
- **241 to 250 GHz.** No known Amateur operation — atmospheric absorption is in excess of 5 dB/km.
- **Above 300 GHz.** This is the highest band (if it can be called that, as there is no upper limit). I know of no Amateur RF work in this region, but if you go up high enough in frequency this "band" includes lightwave (laser) communications, where there is activity.

Many of these bands are shared by radiolocation (radar) services and you must tolerate interference from them. In practice, this isn't a problem. There are also some nearby radio astronomy bands which are protected from Amateur interference.

PHOTO C



Members of the North Texas Microwave Society with their 10-GHz equipment assembled before the first weekend of an ARRL 10-GHz cumulative contest.

How do you get started on the microwave bands? Again, this question has no simple answer; much depends on the individual involved. First, and most important, I'd recommend joining a local group active on microwaves. I can't stress this point too strongly. The group not only provides help and encouragement, but also gives you a number of stations to work. **Photo C** shows members of the North Texas Microwave Society with their collection of 10-GHz equipment, ready for one leg of the ARRL 10-GHz contest. If a local microwave group doesn't exist, try to find a VHF/UHF group. You'll probably find some microwave knowledge and interest there, even if there's no activity. If you can't find either, try to find another ham interested in developing microwave capabilities. You can share knowledge, help each other with construction, and be certain of someone to work once you get your station built! What can you do if you really can't find any help? If all else fails, write to me and I'll try to put you in touch with someone in your area. If I don't know of anyone, I can put your call and address in this column with a plea for help. If that fails, you'll just have to change QTH! It's also important to read up on the subject. There are some good microwave books covering both theory and practice; I'll list a few at the end of this article.

Where should you start to work? I suggest either 23 cm or 3 cm. If you want to work from a fixed station, and you want to find the most activity, 23 cm is the band of choice. While the microwave characteristics of operation on the band make it somewhat different from lower frequencies, much of the equipment and propagation will still seem familiar to VHF/UHF operators. Activity is high enough to make random contacts possible, particularly in urban areas and during activity periods and contests. Most transverter systems use a 144-MHz IF, so a VHF transceiver or transverter is required. Commercial equipment is widely available and a lot of homebrew designs have been published. Check out the "UHF and Microwave Equipment" chapter of a recent *ARRL Handbook* for ideas. Of course it's not necessary to start out with a complete transverter system. Operation using a receive converter and tripler from 432 MHz is quite possible using CW and FM. It's not as convenient as a transverter, but is cheaper, easier to build, and quite capable of yielding equal results. A low power 432 to 1296-MHz tripler can be built using 1N914 diodes (10 cents each!) and, despite a power output of less then 1 watt, I have personally used such a tripler to work distances of >50 km.

On the other hand, if you like the idea of hilltop/portable operation, and want to try out a band with very different characteristics and equipment from the

lower bands, then 3 cm is the band of choice. You won't make any random contacts on this band, so you must be in contact with at least one other interested Amateur. There are two main modes of operation on 10 GHz. First, there are wideband systems with IF bandwidths of 50 kHz or more, which use free running cavity-stabilized Gunn oscillators and usually operate using FM. Second, there are narrowband systems with IF bandwidths of 3 kHz and less, which use frequency multiplication and mixing from lower frequency crystal-controlled sources and operate using SSB and CW. For low cost experimentation, ATV, or digital operation, the wideband Gunnplexer route is best. You can purchase a complete Gunn oscillator-based transceiver from a commercial source for a few hundred dollars. Alternatively, you can build a basic Gunn oscillator/mixer system from a few pieces of waveguide and a couple of diodes for a total cost of around $20. By combining this equipment with a simple power supply and the IF strip from an FM radio (or even an FM radio itself), you can construct a complete transceiver. Some Amateurs are working on ways to convert surplus Gunn oscillator-based microwave radar detectors, automatic door openers, and intruder alarms to Amateur use at low cost. For serious DX and weak signal work under all conditions, narrowband operation is preferred, though it's more expensive and requires a little more microwave knowledge. Narrowband operation will also make possible many more paths than will wideband operation. Wideband systems are quite capable of DX operation under good propagation conditions. (In fact, the world and United States DX records are held by Amateurs using wideband equipment.)

Equipment is available from a number of suppliers. Take a look through this issue of *Ham Radio*; I'm sure you'll find advertisements for companies who specialize in equipment for the microwave bands. All should be able to give you information on microwave equipment, and some may even be able to help you find active microwave stations in your area. If you have trouble finding a specific piece of equipment, I may be able to help. However, it's hard to keep up to date, so check out the ads in the Amateur Radio publications first.

I hope this information is helpful to newcomers who are interested in the microwave bands. There isn't room to print any detailed technical information about how to build simple microwave equipment this month. I hope to return to this topic in the future and try to present some simple projects.

Recommended reading

The following publications are recommended for those who want to learn more about microwaves. Some of these books are available from the HAM RADIO Bookstore.

The RSGB VHF/UHF Manual. Lots of information on theoretical and practical aspects of VHF/UHF and microwave operation. Recommended to anyone interested in these bands. (Available from the HAM RADIO Bookstore for $29.95 plus $3.75 shipping and handling.)

The Gunnplexer Cookbook, by Bob Richardson, published by the Ham Radio Publishing Group. A practical book which describes a large number of projects based on the Microwave Associates Gunnplexer system for 10 GHz. A good start for the newcomer to 10-GHz wideband operation. (Out of print.)

The RSGB Microwave Manual. I still haven't seen this one, but on the basis of its authorship it should be a valuable reference. (Available from the HAM RADIO Bookstore for $35 plus $3.75 shipping and handling.)

10 GHz — A Constructional Project, by Chuck Houghton, published by the San Diego Microwave Group and priced at $15. A collection of notes mostly relating to wideband operation on 10 GHz and some information relevant to narrowband work. Lots of detailed construction information with a little bit of theory. Includes test equipment, use of converted intruder alarms, antennas, homebrew and commercial Gunn oscillators, and more. Some components (IF boards, Gunn diodes, etc.) are also available from this group. Contact Chuck, WB6IGP, at 6345 Badger Lake, San Diego, California 92119.

The RSGB Microwave Newsletter Technical Collection. A collection of technical items from the RSGB microwave newsletter. Covers practical design information for the bands 1296 MHz to 24 GHz. Includes information on oscillator sources, antennas, filter design and test equipment. (Out of print.)

The ARRL also publishes a series of conference proceedings from the Central States VHF Society, Microwave Update, and Mid-Atlantic States VHF Conferences. These publications cover all aspects of operation, theory, and practice on the bands from 50 MHz to lightwave. A good way to keep up with the state-of-the-art technical developments by those in the forefront of VHF/UHF and microwave work. Some, but certainly not all, of the material may be a bit advanced for absolute beginners. (Available from the HAM RADIO Bookstore. Check current book flyer and advertisements in this issue for prices.)

And, of course, back issues of *Ham Radio*. Check out the 5-year cumulative index which appeared in the December 1989 issue. You might also want to read some of the "New Frontier" columns which appeared in *QST* from 1980 to 1989.

Microwave news

As I've mentioned before, microwave operation often takes place on local nets so operators can have a good idea of when and where they'll find activity. WD4MBK has sent along information on a new 1296-MHz net in the Southeast. Dexter McIntyre, WA4ZIA, of Stanfield, North Carolina, has started a net which meets on 1296.090 MHz at 9:30 p.m. every Wednesday evening. The net is held in conjunction with the East Coast 70-cm net which meets on 432.090 MHz at 9:00 p.m., also on Wednesday evenings. On the first night of the 1296-MHz net, there were seven check-ins from five states (Georgia, Florida, Tennessee, South Carolina, and North Carolina). The best DX was a 500-mile contact between W4ODW and WA4ZIA. Stations interested in participating in the 1296-MHz net can check into the earlier 432-MHz net, where net control (WD4MBK) will take a list which will be passed on to the 1296-MHz net controller (WA4ZIA). At 9:30, WA4ZIA will call and listen on 1296.090, while listening simultaneously on 432.110 MHz for stations who wish to join the 1296-MHz net. Stations further to the north (Virginia, Maryland, New Jersey, and Pennsylvania) should look for K4CAW (North Carolina). He will call and listen on 1296.090 at 9:30 p.m. for check-ins from the north.

WA4ZIA has also become operational on 3456 MHz with 5 watts to a W3HQT loop Yagi. To eliminate feedline losses, the 3456 transverter and power amplifier are mounted at the antenna on top of his tower. In his first week on the band, he worked W4OJK

(95 miles) and K4EJQ (115 miles — path open about 10 percent of the time).

Microwaves and no-code

While it's often difficult to be topical in a column written so far in advance of publication, I'm quite sure that the code versus no-code issue will still be under debate when you read this. On February 16th, the FCC issued a notice of proposed rule making (NPRM) in their docket number 90-55. If you haven't read this document, and are concerned with the future of ham radio, please do so. Basically, it calls for dropping the Novice and Technician class licenses and establishing a Communicator no-code license entitling the holder to all privileges on 222 MHz and above. While the debate may be fierce (there are those who argue with a fervor usually reserved for politics and religion), it seems that some kind of no-code license will be the outcome. You may ask what this has to do with microwaves. Well, all the no-code proposals allow — in fact, are actually targeted towards — newcomers to the bands where more activity is badly needed. Much of the fine work done in the UK on the microwave bands has been through the efforts of Amateurs there who hold no-code licenses. (Indeed, my first license was a UK no-code, though my contributions to microwave activity were negligible!) The microwave bands are being eyed increasingly by commercial interests as technology makes their use more practical. If we don't occupy the bands we'll have little grounds for objection if we start to lose them. If a no-code license can increase microwave activity, then I welcome it. Even if it doesn't, I don't think a limited no-code license will do harm.

If you feel strongly on the issue of a no-code license, I urge you to send your comments to the FCC (but do read the text of their proposal first). I favor the idea of the Communicator license, but I think dropping the Novice and Technician classes, with their low speed CW requirements, would be a mistake. Whatever your views, make them known to the FCC, and if the eventual outcome is a new group of no-code Amateurs, then let's make sure we welcome them to ham radio without prejudice and encourage them to join us on the microwave bands (and learn CW!).

Finally, thanks again to all of those who have taken the time to write. The direction and content of this column depends on your letters, so keep them coming. If you send photos, please be sure to include all the information about the photo on the back. Black and white prints are preferred, but color is okay. Next month I plan to discuss tropospheric scatter propagation on the microwave bands.

PHASOMETER

Joel Eschmann, K9MLD, 6964 Meadowdale Drive, Hartford, Wisconsin 53027 and Tom Rehm, K9PIQ, 11653 N. Pinehurst Circle, Mequon, Wisconsin 53092

No, it's not another SWR bridge. The PhasoMeter is a discriminator that detects the relative phase between voltage and current waves in an RF transmission line. When used with a zero-center meter, it provides an accurate indication of resonance and also shows which side of resonance you're tuned to. If you're tuned below resonance, the meter will deflect to the left; if you're tuned above resonance, it will deflect to the right. At resonance, both voltage and current are in phase and the meter reads zero. The further you are from resonance, the greater the amount of deflection. The amount of deflection is also proportional to RF power. The device connects in series with the transmission line or at the antenna feedpoint. You can leave the PhasoMeter "in line" permanently if you wish; there are no batteries to wear out.

Some practical applications include:

- Use as a tuning aid for remote antenna tuners (like the K9MLD design in *Ham Radio*, October 1988[1]) by indicating directly which direction to slew the inductor. This has proven to be invaluable for mobile operation.
- Providing the essential pickup discriminator for construction of an automatic antenna tuner or an audible antenna-tuning indicator for the visually impaired.
- Help in tuning phased arrays to a conjugate impedance match.
- Checking the health of your present antenna. The PhasoMeter is sensitive enough to see your antenna flapping in the breeze — a feat difficult to achieve with an SWR bridge.

Calibration

Calibration requires a dummy load or other flat resistive load. Run 100 to 150 watts through the PhasoMeter into a dummy load. The load end should be the end opposite the point where C4 is attached. Adjust C5 for zero deflection. If you don't obtain a null, move C5 to the other side of T1 and repeat the procedure. That's all there is to it! Calibration should hold to within a few kHz across the entire band.

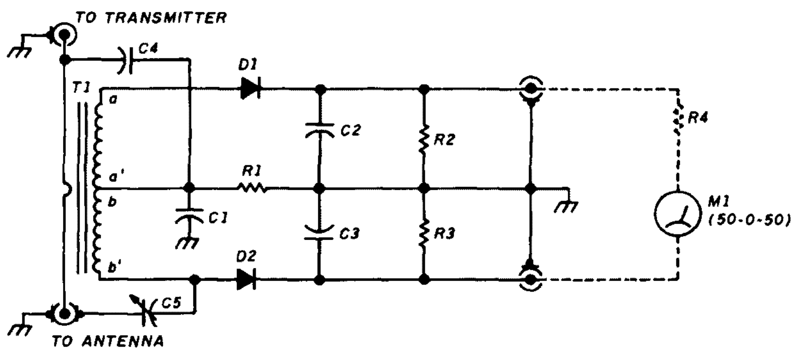
The design in **Figure 1** is optimized for 75-meter operation, with reduced sensitivity on 40 meters. If 40 meters is your band, wind T1 with eight turns.* The basic circuit looks like the early FM detectors, and it is just that. The principle is the same—but now, with the PhasoMeter, you can look into the activity going on in the coaxial cable.

Aids mobile operation

K9MLD has found the PhasoMeter very useful in a mobile

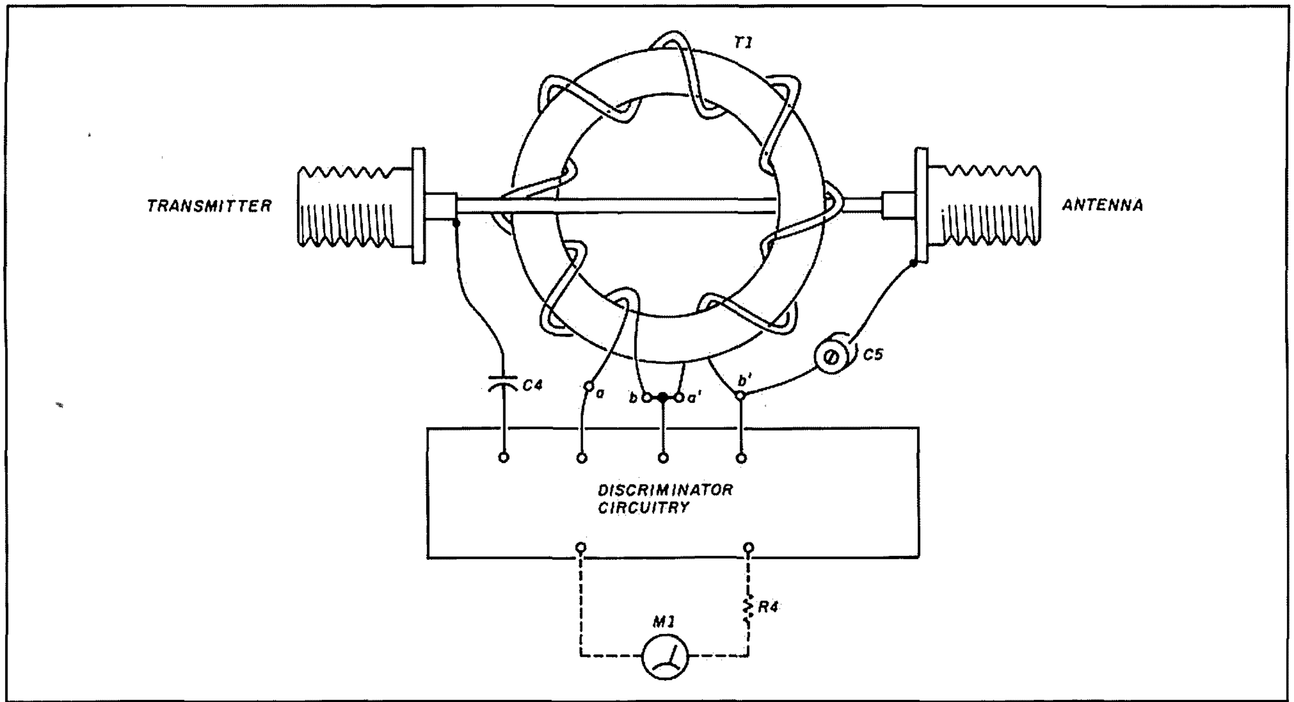
*For optimum performance, T1 should be tailored for the band used. Ed.

FIGURE 1

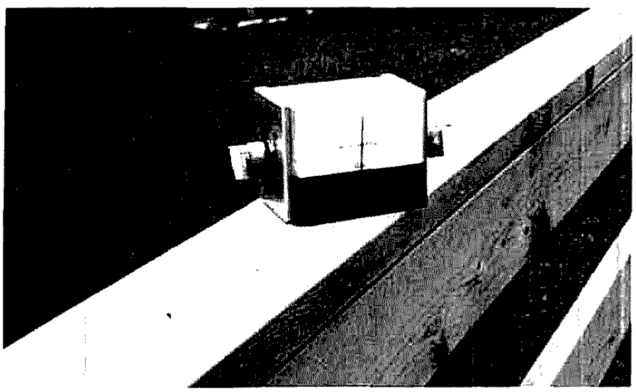


Schematic of PhasoMeter optimized for 75-meter operation, with reduced sensitivity on 40 meters.

FIGURE 2



Pictoral representation of the pickup torold and its windings. Note: For clarity, continuity between the connector bodies isn't shown.



A typical PhasoMeter unit.

PARTS LIST

Parts list

C1	*100-pF silver mica capacitor, 100 volt, 5 percent*
C2,C3	*0.1-μF 100-volt, 5-percent capacitor*
C4	*5-pF silver mica capacitor, 500 volt*
C5	*6 to 50-pF trimmer (Radio Shack)*
D1,D2	*1N914 diode*
M1	*50-μA, zero-center meter*
R1	*12-k, 10-percent, 1/4-watt resistor (a 2.5-mH choke may be substituted for increased sensitivity)*
R2,R3	*100-k, 10-percent, 1/4-watt resistor*
R4	*10-k, 10-percent, 1/4-watt resistor*
T1	*Amidon T-50-2 bifilar wound with 15 turns of no. 26 gauge wire**

*Amidon Associates, 12033 Otsego Street, North Hollywood, California 91607.
PC board available for $3.50 plus $1.50 shipping and handling from FAR Circuits, 18N640 Field Court, Dundee, Illinois, 60118.

installation which uses a manual motor-driven tuner.[1] Changing frequency and retuning the antenna while mobile can be a real chore. This unit makes it simple. The zero-center meter clearly indicates the direction to tune to obtain resonance.

Figure 2 shows how to assemble the unit. The toroidal core and detection circuitry should be located within an inch or so of each other. The meter can be remote to your dashboard or the whole unit can be mounted next to the transmitter. We've built a number of these meters and found the performance to be equal, regardless of the structure. A typical unit is shown in Photo A.

Future plans

We are working on an addition to the PhasoMeter that will automatically drive a motor-driven tuner to resonance. Though this type of design has been around a while, the earlier models used a "bang bang" style of servo. The new design will servo the motor and slew into the point where the motor is at rest. When the motor-driven reactor approaches resonance, our design will slow the motor down, come into position at the proper point, and stop — instead of stopping abruptly and over-shooting the mark.

We have described a discriminator-type device that attaches to your transmission line and has many potential uses. You now have a tool that will make tuning your antenna system even easier and keep it operating at peak performance.

REFERENCES

1. Joel Eschmann, K9MLD, "The Weekender: Remote Tuner For 75-Meter Mobiles," *Ham Radio*, October 1988, page 36

INEXPENSIVE SSB FOR 10 METERS

Modify this CB radio for your 10-meter needs

By Phil Salas, AD5X, 1517 Creekside Drive, Richardson, Texas 75081

Ten meters is becoming an exciting place again. The sunspot cycle is heading for good times over the next few years, and now Novices and Technicians can operate phone over a limited portion of this band (28.3 to 28.5 MHz). There's an inexpensive way for everyone to get on board. It'll give you a chance to dust off the old soldering iron and get your own rig on the air.

CB radios are still very reasonable. Though the SSB rigs are a little more costly, they're still a good bargain. I purchased a brand-new Cobra 146GTL from Fordham for about $150. Converting this unit to the 10-meter ham band is a relatively simple 1 to 2-hour job. The results are well worth it.

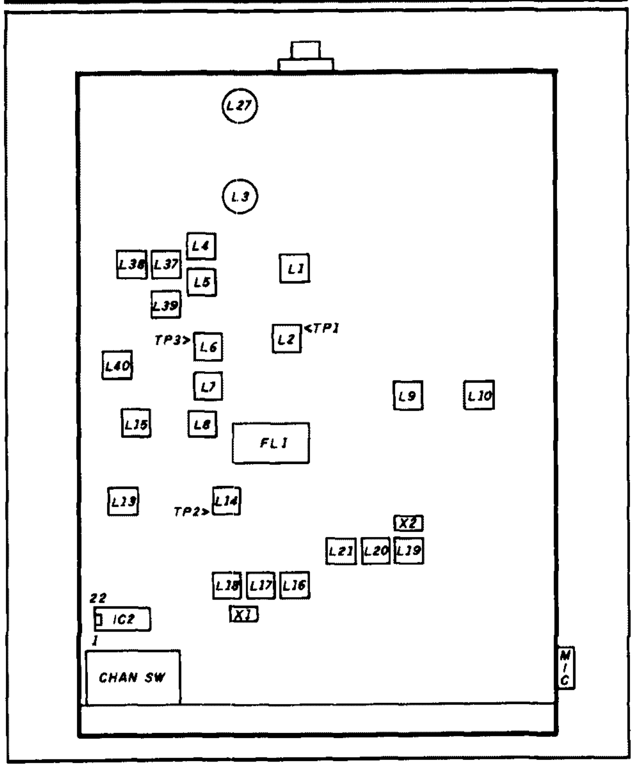
The Cobra 146GTL is a fully synthesized 40-channel AM and SSB radio. It has an RF gain control, automatic noise limiting, and noise blanking. The receiver is robust and doesn't suffer overload from nearby CB radios. The SSB filter is an eight-pole centered at 10.695 MHz. The transmitter puts out 12 watts PEP on SSB and 4 watts on AM.

There are three main areas of modification. The first involves changing the receiver voice lock control circuitry to permit continuous tuning of both transmitter and receiver. Next you'll modify the synthesizer circuitry to permit operation on 10 meters. Your final step will be to retune the transmitter and receiver for optimum performance on the 10-meter band. Before you start, remove the top and bottom covers. You might also want to extend the wires going to the speaker. You'll be turning the unit over quite a bit, and the original speaker wires are too short. Refer to **Figure 1** for parts placement and become familiar with their locations. Find these same parts on the schematic diagram supplied with your 146GTL. Make sure your unit is in working condition before attempting these modifications.

Voice lock modification

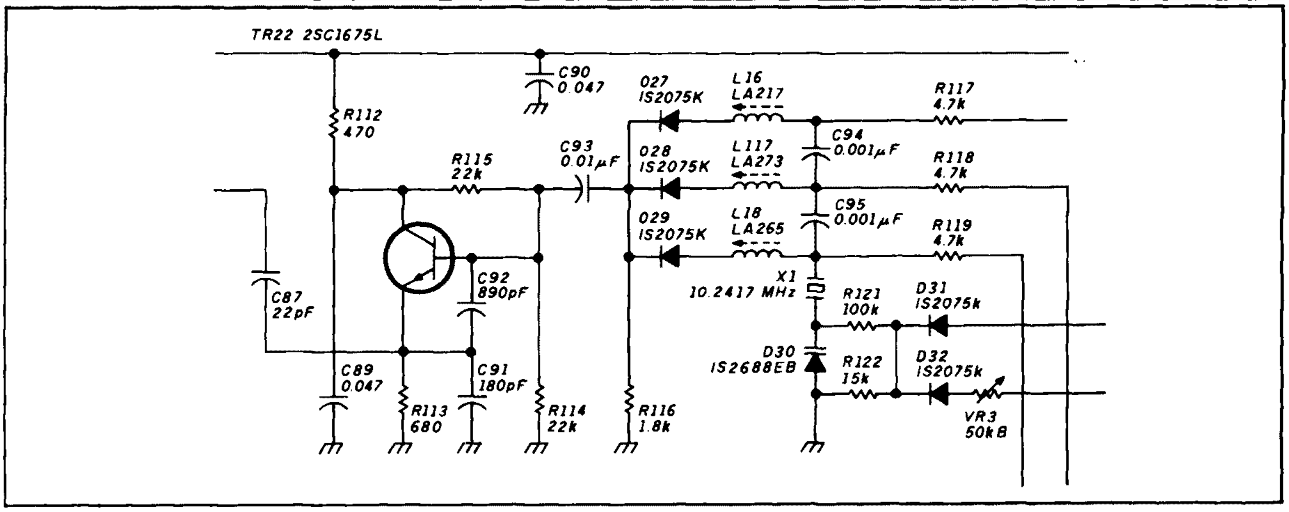
CB radios only permit transmit operation on what are basically 10-kHz fixed steps. When you're receiving there's usually a receive voice lock or clarifier control that permits a ±1 kHz tuning range. This is necessary because of slight differences in frequency accuracy between radios. Though it's not really necessary for AM operation, you must be tuned precisely to the received signal when operating SSB to prevent severe audio distortion. The trick is to adjust the voice lock circuitry so that it has enough range to cover the whole 10-kHz channel and also work on both transmit and receive.

FIGURE 1



Parts location layout for the Cobra 146GTL.

FIGURE 2



Partial schematic for the reference oscillator in the Cobra 146GTL.

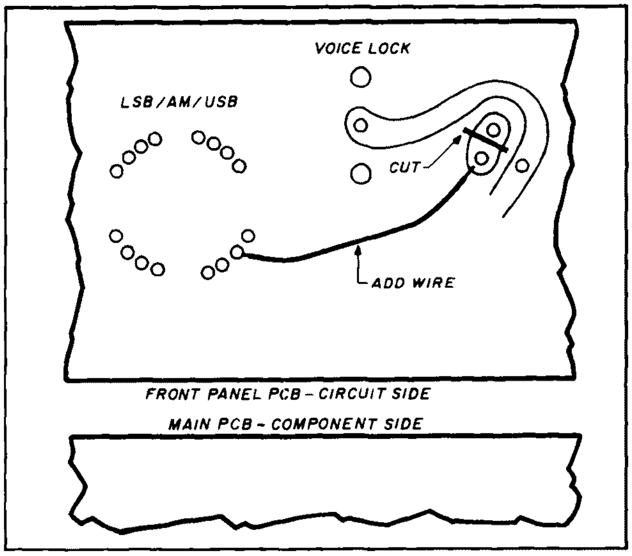
Figure 2 is a schematic of the reference oscillator for the synthesizer in the 146GTL. Diode D30 is a varactor diode (voltage variable capacitor). On receive the voice lock control applies a variable voltage through isolation diode D31 to varactor diode D30. This variable voltage changes the capacitance of D30 which then pulls the frequency of synthesizer crystal X1. On transmit the voice lock voltage is removed and a fixed voltage is applied through isolation diode D32 to varactor diode D30.

First disable the fixed voltage applied during transmit. This is done most easily by cutting D32 out of the circuit. D32 is located just beneath crystal X1. Next refer to **Figure 3** which is a partial drawing of the circuit side of the front panel pc board. Modify this board by cutting the circuit pad near the voice lock control and adding a wire from the cut pad to one terminal of the LSB/AM/USB switch. This modification will keep the voice lock variable voltage available during receive and transmit.

As I mentioned before, the total tuning range of the voice lock circuitry is only ±1 kHz. Varactor diode D30 doesn't have enough capacitance range to cover any more than this. Fortunately, you can make a very inexpensive varactor diode. Simply replace D30 with four 1N4000 diodes connected in series as shown in **Figure 4**. Note that the diode polarities are opposite that of D30. The voice lock circuitry now becomes a current source for these diodes. At very low currents the diodes don't conduct, and look like an open circuit or very low capacitance. As you increase the current through the diodes, they begin to conduct. The more the diodes conduct, the more they approach a short circuit. Remember that a short circuit looks like a very high capacity. I found that three series-connected diodes gave me ±5 kHz tuning range while four series-connected diodes gave me a ±7.5 kHz tuning range. Four diodes makes tuning a little more sensitive but gives you some channel overlap.

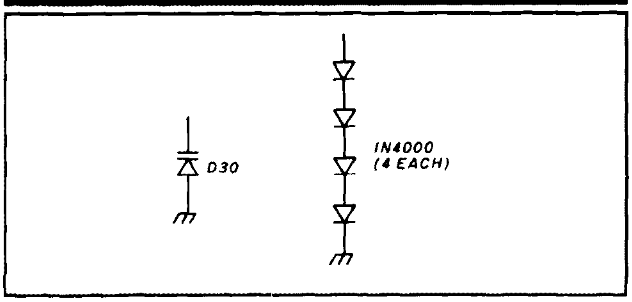
The Cobra 146GTL uses crystal X1 to determine the USB, LSB, and AM frequencies. To keep the frequencies the

FIGURE 3



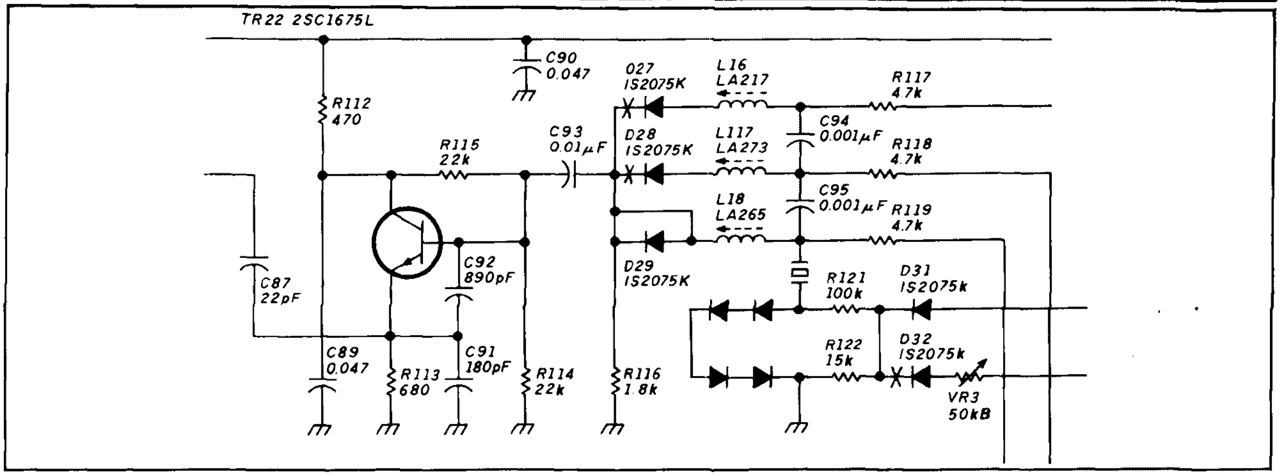
Partial drawing of the trace side of the front panel circuit board.

FIGURE 4



Replacing D30 with four 1N4000 series diodes.

FIGURE 5



Schematic of the modified reference oscillator.

same for all three modes, the crystal is pulled by inductors L16, L17, and L18 — depending on which mode you're in. Unfortunately, the combination of the inductor pulling and the voice lock pulling is a little too much for the crystal oscillator circuit in the USB and AM modes. Consequently, the oscillator may stop while you're tuning with the voice lock control. To compensate, I modified the circuit to use inductor L18 (the LSB inductor) for all three modes. Do this by cutting one end of diodes D27 and D28 and replacing diode D29 with a short circuit (a piece of wire). You can still operate all three modes, but there will be a slight frequency change when you switch modes.

Your final hardware modification involves replacing crystal X1. An 11.1-MHz crystal will move the operation of the 146GTL to 28.3 to 28.74 MHz. For each 100 kHz you move the crystal frequency, the radio operating frequency will move 200 kHz. For example, if you want the radio to operate from 28.5 to 28.94 Mhz, use a crystal frequency of 11.2 MHz. I ordered my replacement crystal from JAN Crystals for $5. When you order the crystal, specify that it must be a synthesizer crystal for the Cobra 146GTL in an HC-18 package with a frequency of 11.100 Mhz. When your crystal arrives, unsolder crystal X1 and replace it with the new one. **Figure 5** is the schematic of the final modified synthesizer reference oscillator.

Tuning up

Two of the necessary adjustable inductors (L14 and L27) in the Cobra 146GTL are filled with wax. You can remove this wax easily with a small bladed X-ACTO™ knife. I was able to remove virtually all the wax in just a few minutes by picking at it with the knife.

It's best to perform the following adjustments with a digital voltmeter and an oscilloscope. If you use an analog voltmeter, make sure that it is high impedance. You can use a CB-style power/SWR meter in place of the oscilloscope for the final output power tuning; however, the oscilloscope is really necessary for setting up the synthesizer circuitry.

Make sure the microphone is plugged into the radio. The receiver won't function without the mic.

Proceed as follows:

Make sure the crystal oscillator is operating. You should see about 50 mV p-p of signal with the oscilloscope connected to pin 13 of IC2. This is a broadband circuit and will operate easily with the new 11.1-MHz crystal. If it's not working, recheck your modifications.

Set the channel selector to channel 19. Locate TP2. This is the bare end of resistor R93 and is located next to L14 (see **Figure** 1). Attach your digital voltmeter between TP2 and the 12-volt return. Carefully adjust L14 in a counterclockwise direction (for less inductance) until you read 3.25 volts at TP2. As you begin to adjust this inductor, the voltage will start out low (less than 2 volts) and then increase. When you reach 3.25 volts, connect the oscilloscope to pin 22 of IC2 and peak the observed voltage with L13. You should see aproximately 130 mV p-p at this point.

If the voltage at TP2 becomes intermittent while you're tuning L14, leave the inductor at the last stable voltage position and peak the voltage at pin 22 of IC2 with L13. Then go back and finish readjusting L14. Follow this by peaking the observed voltage with L13 at pin 22 of IC2.

Now connect the oscilloscope to TP3. TP3 is one end of resistor R105 and is located next to L6 as shown in **Figure 1**. Adjust L15 for maximum voltage at TP3. *Note: Adjust L15, not L6!* You should see about 80 mV p-p at TP3.

Connect the oscilloscope to pin 11 of IC5. Press the transmit button and adjust L39 for maximum voltage. This should be about 50 mV p-p.

Finally, connect the oscilloscope to the center conductor of the output connector on the radio. With the transmit button depressed, adjust L38, L37, and L27 for maximum output. This completes the transmitter alignment.

The receiver alignment is also simple. Attach your oscilloscope probe to one of the speaker leads. Set the channel selector to channel 19 and attach a 10-foot section of random wire to the radio's antenna terminal. Tune around for a clear area and then peak L3, L4, L5, L6, L7, and L8 for maximum noise as seen on the oscilloscope. Make sure you don't press the transmit button since you don't have a proper antenna!

Attach your oscilloscope probe to TP1. This is one end of resistor R10 and is located adjacent to L2, as shown in **Figure** 1. Now adjust L1 and L2 for maximum noise. This completes the radio alignment.

Antennas

Any 10-meter antenna will work. But for mobile use I wanted a discreet antenna that I could remove when I parked the car. I purchased a center-loaded CB magnetic mount antenna from K-Mart for about $15. I removed 1-3/4 inches from the top part of this antenna to resonate it at 28.5 MHz. The entire antenna is only about 2 feet high!

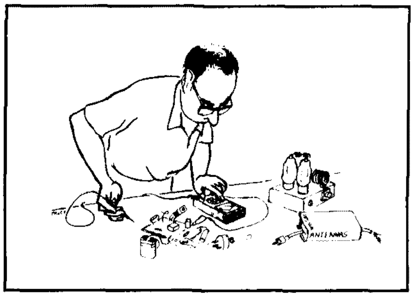
Operation

What can you do with a 2-foot antenna and 12 watts PEP on 10 meters? It depends on band conditions. I work all over the United States and down into South America quite regularly. If I can hear a station I can work it, unless I'm competing with someone running higher power and a better antenna system. My best results seem to occur when I call CQ and state that I am "QRP mobile." My contacts are always amazed when I tell them what I'm running.

Summary

It's hard to beat 10 meters during the sunspot highs. Over the next few years you can have quite a bit of fun with this rig, for almost half what you'd spend for a 2-meter handie-talkie! The modifications I've described can be done easily in one evening. Give it a try. If you're like me, you'll wish you had a longer drive to work!

Ham Radio Techniques



By Bill Orr, W6SAI

WARC 1992

In 1992, if all goes as planned, a World Administrative Radio Conference (WARC) will convene in Geneva, Switzerland. The Amateur Radio bands will be examined closely once again by the participants, some of whom view the Amateur frequencies as a happy hunting ground. We were lucky at the last WARC. We retained our traditional HF bands *and gained new narrow bands* at 10, 17, and 24 MHz. Good long range planning by the United States delegates and observers, in cooperation with delegates from other countries, brought us this triumph.

Now we must run the gauntlet of a new conference held during perplexing and rapidly changing times. Our technical and political worlds are in flux, and this situation raises some important questions. In view of the amazing changes in Eastern Europe, are the costly, spectrum-consuming shortwave broadcasts of special interest stations like Radio Free Europe and Radio Liberty still necessary? Are the propaganda broadcasts of Radio Moscow, Radio "Free" Cuba, and numerous other countries relevant?

Sadly, the "radio" portion of the cold war has been growing instead of shrinking. Many countries are, or will soon be, on the air with megawatt HF broadcast transmitters operating into huge curtain arrays. Transmissions are made around the clock in many foreign languages. Radio Peking, for example, broadcasts to the world in over 40 languages including Serbo-Croatian, Urdu, and Bulgarian. I wonder what common bond the speakers of these languages have with China that would induce them to listen to these broadcasts? It seems to me that many countries squander megawatts of valuable energy broadcasting their views to the world.

The upshot of all this is that the international shortwave broadcast bands are a mess. The stations are layers deep! It's interesting to contemplate the return on the investment of these broadcasts. *It must be very low.* Regardless, many countries (including the United States) are upgrading their shortwave facilities with more power, bigger antennas, and better studios.

No doubt there will be enormous pressure at WARC '92 for an expansion of the present shortwave broadcast bands to accommodate all the new voices and provide more elbow room. I think they should do the reverse. Let's clean up the present bands by cutting down on repetitive broadcasts! Eliminate multiple transmissions of the same program in a particular band. Place a cap on the power war. Treat the bands as a valuable resource instead of a propaganda dumping ground.

During the cold war decades, the world grew tired of propaganda. Most people aren't fooled by what they hear on the shortwave radio. For years, casual listeners were amused by the "news" broadcasts from Radio Moscow. Megawatts of energy, large transmitting facilities, and a huge staff were committed to these broadcasts. They went on 24 hours a day, 365 days a year, blanketing the world with bombast in every known language. Any educated listener knew he was hearing hogwash. For years the Pacific Service of Radio Moscow bruised the ears of listeners in the Americas with broadcasts of such poor audio quality and high distortion that the programming was virtually unintelligible. Yet the broadcasts went on. It seemed the desire was to transmit something — anything — whether the transmission was readable or not.

It would be helpful if the forthcoming WARC were to examine some of the gross abuses occurring in the HF broadcast spectrum. If these bands were being used properly, there would be plenty of room for all. Spectrum conservation would help to take the heat off the Amateur bands.

From a technical viewpoint, the same can be said of the fixed service bands. More and more communicators are using satellite transmission. In the meantime, time-honored marine CW stations (like Radio San Francisco, KFS) are on the verge of closing down. SSB has supplanted AM; SITOR is rapidly replacing FSK and CW. This means there's more space available in the fixed service and maritime bands. And, if all goes well, the military requirement for HF radio should decrease as the armed forces of the world shrink!

Great opportunities for Amateur Radio lie ahead if our representatives prepare themselves beforehand and work in harmonious conjunction with Amateur Radio societies around the world to establish a strong and realistic position. Of course, this means we must get our own house in order!

Yes, the next year or so will certainly be interesting as far as the future of Amateur Radio is concerned. Stay tuned.

The ON4UN Yagi design program

Computers can be a great help in determining the electrical and physical properties of an antenna. In past columns I've discussed the antenna programs available from Brian Beezley, K6STI. These are user friendly derivatives of MININEC, a program developed at the Naval Ocean Systems Center at Point Loma, California. The K6STI programs provide modeling and optimization facilities that are very helpful to the antenna designer. They also include a small library of predesigned antennas.

John Devoldere, ON4UN, and Roger Vermet, ON6WU, have compiled a software package which takes you through all the aspects of Yagi design

FIGURE 1

DESIGN # 10 ELEMENTS: 3 NAME: FREDA BOOM: 0.249 WVL

FREQ.	GAIN	F/B	RESIST	REACT.	SWR	FOM	
-1.5%	7.4	20.5	28.8	-12.6	1.5	9.0	ANT. Q = 14
-1.0%	7.4	23.4	29.1	-9.0	1.3	9.4	
-0.5%	7.4	23.9	28.8	-5.3	1.1	9.7	SWR BW > 3 %
0.0%	7.5	24.4	28.0	-1.4	1.0	9.9	
+0.5%	7.5	25.4	26.8	2.8	1.2	9.9	F/B BW > 3 %
+1.0%	7.6	22.8	25.2	7.5	1.4	9.5	
+1.5%	7.7	19.2	23.4	12.5	1.8	8.9	

ELEMENT	LENGTH	POSITION
REFLECTOR	0.510217	-.105595
DRIV. EL.	0.483032	0.000000
DIR # 1	0.452359	0.143370
DIR # 2	0.000000	0.000000
DIR # 3	0.000000	0.000000
DIR # 4	0.000000	0.000000

PHYSICAL BOOMLENGTH		
28 Mhz ->	2.63 m. OR	8.6 ft
24 Mhz ->	2.99 m. OR	9.8 ft
21 Mhz ->	3.52 m. OR	11.5 ft
18 Mhz ->	4.12 m. OR	13.5 ft
14 Mhz ->	5.27 m. OR	17.3 ft
10 Mhz ->	7.37 m. OR	24.2 ft
7 Mhz ->	10.51 m. OR	34.5 ft

EL. LENGTHS ARE FOR EL. DIAM. OF .0010527 WAVELENGTHS (7/8 INCH ON 14.2 MHz).
S = SELECT THIS DESIGN C = CONTINUE M = MENU

The "generic" design for a three-element beam. Length and position ot the elements are given in wavelengths. Gain (dBi), front-to-back ratio, and input impedance are provided over a band of frequencies.

— electrical as well as mechanical. This is not strictly a modeling program; it consists of a comprehensive database of 100 different HF Yagis from two to six elements. There are Yagis for every application: wideband, narrowband, gain optimized, front to back optimized, and so on.

Working from the program's main menu, you can browse through the database and select a Yagi design that meets your needs. You can specify maximum boom length, minimum gain, or other parameters. You can modify the design as you see fit. As an aid, a well-written program outline is provided on the disk.

The Figure of Merit

In addition to the usual parameters, the ON4UN program expresses each Yagi design in terms of a Figure of Merit (FOM) which combines the main performance of a Yagi in one figure:

FOM = gain (dBi) + 0.1 × (worst lobe front to back) − (1 − SWR)

The Figure of Merit is an ideal tool for rapidly evaluating the overall (gain, F/B, and SWR) bandwidth performance of a Yagi. Armed with the FOM of the various antennas in the ON4UN database, you can quickly find an antenna that will fill your needs.

If you want to add your own design to the existing database, the program provides an empty database with room for up to 100 records. That should satisfy even the most unusual antenna requirements!

Once you've selected a design, the program provides the dimensions for untapered ("generic") elements (see **Figure 1**). The program also provides a tapering schedule and matching data for gamma, omega, and hairpin matches.

Mechanical design of the Yagi

A novel and valuable section of the ON4UN program concerns Yagi mechanical aspects. Element strength, boom strength, weight and balance, wind loading and torque moment can be determined easily for a given design. All of these characteristics are illustrated in the program for Design no. 10: a *three-element 40-meter* beam which provides better than 7-dBi gain and 20 dB or more F/B ratio over the entire band. The boom length is only 24.5 feet! SWR across the 40-meter band is less than 1.8:1.

The ON4UN database

It's both instructive and entertaining to browse through the ON4UN program database. Each antenna has been given a woman's name. I like CYBIL (Design no. 15) for 20 meters. It offers a very good F/B ratio and excellent gain. The antenna fits on a 20-foot boom and can be considered a classic design which can be scaled to other bands (see **Figure 2**).

The 10-meter contest operator may wish to choose JOAN (Design no. 95), a wide-spaced (1.03 wavelength), six-element stagger-tuned beam. This antenna provides 11.6-dBi gain with an excellent F/B ratio. It covers the range 28.0 to 28.75 with less than 2:1 SWR. Not bad!

Summary

All in all, the ON4UN program is a valuable tool for the Amateur interested in antenna performance. The YAGI DESIGN software comes in MS-DOS format. For more information contact *either John Devoldere, Poelstraat* 215, B9220 Merelbeke, Belgium; B.W. Jorden, K7KI, 6861 Kenanna Place, Tucson, Arizona 85704; or HAM RADIO Bookstore, Greenville, New Hampshire 03048-0498.

FIGURE 2

DESIGN # 15 ELEMENTS: 3 NAME: CYBIL BOOM: 0.288 WVL

FREQ.	GAIN	F/B	RESIST	REACT.	SWR	FOM		
-1.5%	7.4	23.8	36.0	-15.3	1.6	9.2	ANT. Q =	18
-1.0%	7.5	24.5	34.1	-11.2	1.4	9.5		
-0.5%	7.6	25.2	31.8	-6.7	1.2	9.9	SWR BW >	3 %
0.0%	7.7	26.1	29.3	-1.6	1.0	10.3		
+0.5%	7.8	21.9	26.6	4.0	1.3	9.7	F/B BW >	3 %
+1.0%	7.9	18.4	23.9	10.1	1.6	9.1		
+1.5%	8.0	15.6	21.2	16.9	2.2	8.4		

ELEMENT	LENGTH	POSITION
REFLECTOR	0.512185	-.156638
DRIV. EL.	0.485799	0.000000
DIR # 1	0.458877	0.131112
DIR # 2	0.000000	0.000000
DIR # 3	0.000000	0.000000
DIR # 4	0.000000	0.000000

PHYSICAL BOOMLENGTH				
28 Mhz ->	3.04 m.	OR	10.0 ft	
24 Mhz ->	3.46 m.	OR	11.3 ft	
21 Mhz ->	4.07 m.	OR	13.3 ft	
18 Mhz ->	4.76 m.	OR	15.6 ft	
14 Mhz ->	6.09 m.	OR	20.0 ft	
10 Mhz ->	8.52 m.	OR	28.0 ft	
7 Mhz ->	12.15 m.	OR	39.9 ft	

EL. LENGTHS ARE FOR EL. DIAM. OF .0010527 WAVELENGTHS (7/8 INCH ON 14.2 MHz).

S = SELECT THIS DESIGN C = CONTINUE M = MENU

Three-element Yagi for 20 meters. Boom length is 20 feet. Design may be scaled to other bands by program.

FIGURE 3

DESIGN # 14 ELEMENTS: 3 NAME: CHRISTY BOOM: 0.288 WVL

FREQ.	GAIN	F/B	RESIST	REACT.	SWR	FOM		
-1.5%	7.2	21.8	44.6	-13.3	1.4	9.0	ANT. Q =	11
-1.0%	7.2	22.6	42.9	-10.0	1.3	9.2		
-0.5%	7.2	23.1	30.7	-6.2	1.3	9.3	SWR BW >	3 %
0.0%	7.3	23.7	38.3	-2.0	1.0	9.7		
+0.5%	7.4	24.5	35.5	2.6	1.2	9.7	F/B BW >	3 %
+1.0%	7.5	25.4	32.6	7.8	1.4	9.6		
+1.5%	7.6	21.6	29.6	13.6	1.7	9.1		

ELEMENT	LENGTH	POSITION
REFLECTOR	0.514716	-.165057
DRIV. EL.	0.487664	0.000000
DIR # 1	0.454808	0.122693
DIR # 2	0.000000	0.000000
DIR # 3	0.000000	0.000000
DIR # 4	0.000000	0.000000

PHYSICAL BOOMLENGTH				
28 Mhz ->	3.04 m.	OR	10.0 ft	
24 Mhz ->	3.46 m.	OR	11.3 ft	
21 Mhz ->	4.07 m.	OR	13.3 ft	
18 Mhz ->	4.76 m.	OR	15.6 ft	
14 Mhz ->	6.09 m.	OR	20.0 ft	
10 Mhz ->	8.52 m.	OR	28.0 ft	
7 Mhz ->	12.15 m.	OR	39.9 ft	

EL. LENGTHS ARE FOR EL. DIAM. OF .0010527 WAVELENGTHS (7/8 INCH ON 14.2 MHz).

S = SELECT THIS DESIGN C = CONTINUE M = MENU

Design no. 14 covers lower half of 10-meter band with low SWR (see *Figure* 6).

FIGURE 4

```
DESIGN # 14                NAME: CHRISTY                            ELEMENTS =  3
---------------------------------------------------------------------------------
FREQ. =  28.5  MHz         WAVEL.:  10.5193 m.      BOOM:  3.03 m or    9.93 ft
DRIVEN ELEMENT REACTANCE= -2  ohm.
---------------------------------------------------------------------------------
ELEMENT/POSITION             CENTIMETERS      INCHES      WAVELENGTHS
----------------             -----------      ------      -----------
POSITION REFLECTOR             -173.6          -68.4        -.165057
LENGTH REFLECTOR                544.4          214.3        0.517526
POSITION DRIV. ELEM.              0.0            0.0        0.000000
LENGTH DRIV. ELEM.              511.1          201.2        0.485876
POSITION DIR # 1                129.1           50.8        0.122693
LENGTH DIR # 1                  471.5          185.6        0.448200

THESE LENGTHS ARE FOR A CONSTANT DIAMETER OF 7/8 INCH OR 2.2225 CM.

 1=MENU    2=FREQ.CHANGE    3=TAPER    4=MATCH    5=DESIGN    6=CHANGE DR. EL
```

Program provides element length and boom position. Next step is to determine effect of tapered element (about 1 inch).

A three-element Yagi for 10-meter DX

Last month's column featured a wideband Yagi that covered the whole 10-meter band. It provided moderate gain and good front-to-back ratio. For the serious DXer, the lower portion of the band holds the greatest interest and Design no. 14 (CHRISTY) in the ON4UN program provides an ideal Yagi beam (**Figure 3**) for this type of operation. Gain is 7.3 dBi with an F/B ratio of 38 dB at the design frequency of 28.5 MHz. The design requires a 10-foot boom (**Figure 4**).

Elements are mounted above the boom by means of a flat plate measuring 6 × 2 inches; U bolts hold the element to the plate and the plate to the boom. The elements are made of 7/8 and 3/4-inch tubing with a wall thickness of 0.058 inch. The elements are telescoping. The slight degree of taper increases the element length 1 inch from the generic design.

A gamma matching section is used, as shown in **Figure 5**. The gamma is made of an aluminum tube. The center conductor of RG-8/U is inserted into the tube to form the gamma capacitor.

The match is excellent (see **Figure 6**). SWR at 28 MHz is 1.35, at 28.5 MHz it's unity, and at 28.9 MHz it's 1.65. The F/B ratio is better than 21 dB over this frequency range, peaking at 38 dB at the design frequency.

You say you'd like a bigger beam with more gain? Just look in the ON4UN library and make your choice!

FIGURE 5

```
                         GAMMA MATCH DESIGN
                         ------------------
INPUT DATA
----------
DESIGN FREQUENCY :  28.5 MHz.
FEEDLINE IMPEDANCE :  50 ohm.
ANTENNA POWER :  1200 WATT.
Z-ANT RESISTIVE PART :  38.30  ohm.
Z-ANT REACTIVE PART :  -2.00 ohm.
ELEMENT DIAMETER :  .875
GAMMA ROD DIAMETER :  .375                   STEP UP RATIO =  5.91
SPACING (CENTER TO CENTER) :  3

RESULTS
-------
GAMMA ROD LENGTH :  54.0 cm  OR  21.3 inch.
SERIES CAPACITOR :  99 pF.
VOLTAGE ACROSS SERIES CAPACITOR :  276 Volts.
CURRENT THROUGH SERIES CAPACITOR :  4.9 Amp.
```

After taper is determined, program provides data for gamma match, including voltage and current rating of series capacitor.

Errata

In my March column, my post office box was given as 7805. The correct address is Box 7508, Menlo Park, California 94025.

Also in March, I stated that the radiation resistance of a folded Marconi was four times that of a single-wire Marconi. This is incorrect; the radiation resistance of the two antennas is the same. The feedpoint impedance of the folded Marconi, however, is four times that of the single wire equivalent.

Thanks to the many sharp-eyed readers who caught these goofs!

The Dead Band Quiz

K4COF proposed the following: Can the three hands of a conventional analog clock ever trisect the circle? In a nutshell, for an ideal analog clock (that is, one where each hand moves continuously and uniformly), the answer is no.

If the hands are detented, KAØPGA points out that two "obvious" solutions are 4:00:40 and 8:00:20. WD8KBW follows up with the statement that there are 22 non-integer solutions; WØNI says one of these is 9:05:2524...hours with only a 0.172-degree error in the second hand.

FIGURE 6

DESIGN # 14 GAMMA MATCH

FREQ. DIFF (%)	-1.5%	-1.0%	-0.5%	0.0%	+0.5%	+1.0%	+1.5%
FREQUENCY	28.073	28.215	28.358	28.5	28.643	28.785	28.928
R(ANT)	44.6	42.9	30.7	38.3	35.5	32.6	29.6
X(ANT)	-13.30	-10.00	-6.20	-2.00	2.60	7.80	13.60
R(OUT)	57.90	56.40	58.50	50.00	44.80	38.30	31.30
X(OUT)	13.20	9.10	-10.80	0.00	-3.90	-6.40	-7.20
SWR	1.33	1.23	1.29	1.00	1.15	1.35	1.65

Program provides gamma match specifics at seven frequencies, plus antenna impedance data.

AE2P and George McHugh indicate that 2:54:33 and 9:05:27 are close. KM4AS wonders about 11:38:10.9090... and 12:21:49.0909....

KØYXE, KE6VK, and others sent a BASIC program which lists all the permutations and combinations of clock hands and the error in angularity of a large number of approximate clock values that come close, but aren't exact.

I can't thank each reader who sent in solutions and comments personally, but I appreciate the time and effort put into all the replies. I read and enjoyed each one!

To date, the following have responded to this quiz: Jay Harvey, KAØPGA; Paul Bunnell, KE6VK; Stan Kadron, W4UGW; George McHugh; Franklin Antonio, N6NKF; John Fowler, AA5HR; Les Moskowitz, AE2P; Dave Roberts, WØNT; Paul Lalli, AA5AN; Irene Kott, WO8E; Wayne Cooper, AG4R; Dan Hopper, K9WEK; Mike McDermott; Ron Romer, N1BHE; Bill Shanney, KJ6GR; Bob McGraw, W2LYH; Art Lashbrook, WX6L; Al Weller, WD8KBW; Tim Bratton, K5RA; John Bellah, KØYXE; Bryan Suits, WB8WKN; Don Miller, KM4AS; and Ted Kroener, KA1PL.

VARACTOR DIODES FROM RECTIFIERS

A new application for rectifier diodes

Hugh Wells, W6WTU, 1411 18th Street, Manhattan Beach, California 90266

Many circuit applications, like frequency synthesizers and RF resonators, require varactor diodes for tuning. The name varactor comes from a combination of words like voltage-variable capacitor and voltage-reactance diode. Varactor diodes, also called varicaps, provide remote tuning capability through a voltage change as opposed to the mechanical movement of a variable capacitor.

Unfortunately, varactor diodes aren't always easy to find. But it occurred to me that rectifier diodes might work as substitutes. To find out, I set up an experiment to classify various diodes for capacitance as a function of applied voltage.

How they work

When a diode is made, a conduction barrier is formed at the junction of its P and N materials. This barrier is called the depletion layer. Applying voltage to the junction causes the barrier to narrow or widen, depending upon the polarity and magnitude of the applied voltage. This voltage application is referred to as forward and reverse bias. With forward bias the barrier narrows, allowing conduction to take place. When the bias is reversed, the barrier widens and serves as an insulator. The percentage of barrier movement under voltage control is predominately a function of material doping. Rectifier diodes are heavily doped to enhance a high forward current; this results in small reverse-biased barrier width changes.

Widening the barrier produces an effect which is similar to that achieved by increasing the spacing between plates of a capacitor — the capacitance value is reduced. Reverse-biased diodes exhibit a decrease in capacitance between their terminals as the applied voltage is increased.

Experimental results

Because rectifier diodes weren't designed for varactor applications, the capacitance value tends to vary somewhat from one device of a given type to another. But sufficient similarity exists to satisfy most Amateur applications. As expected, point contact diodes exhibited a lower value of capacitance than diffused junction diodes. Point contact diodes also have a smaller percentage of capacitance value change for a given applied voltage range. **Table 1** shows the values obtained from samples of various diodes. I observed that a minimum voltage from 0.5 to 1 volt was required for the diode to start acting like a capacitor. Below the minimum voltage value the diode dissipation was high, but decreased as the voltage was increased. As the applied voltage was increased, a value was reached where the capacitance stopped decreasing, as if saturation had occurred. I met the objective for characterization by limiting the experiment to a maximum of 16 volts. However, a higher voltage might have provided some additional data. The voltage range obtained for controlling the capacitance change seemed to be independent of the peak reverse voltage (PRV) value of the diode.

Measurement method

Measuring low picofarad values of capacitance with accuracy is difficult, particularily when the capacitance value being measured is nearly equal to the stray circuit capacitance value. I attached a power supply to the diode to provide the DC bias needed for varying the capacitance. My objective was to identify obtainable capacitance values and range as a function of the applied voltage.

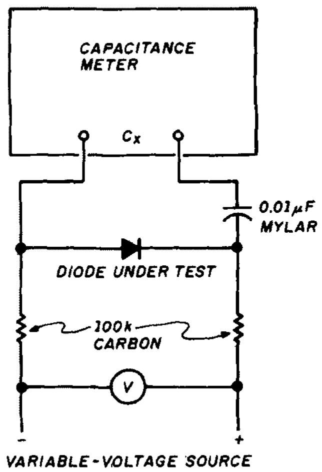
I used both digital and analog capacitance meters to

TABLE 1

Diode capacitance values in pF.

	Voltage, volts							
	0.5	1	2	3	4	5	10	16
Diode	Capacitance, pF							
1N34	4	3.5	2	-	-	0.5	-	-
1N92	170	160	140	120	90	70	19	13
1N295	3	2	1	0.5	0.5	-	-	-
1N305	-	5	4	3	2.9	2.8	2	2
1N457	13	11	7	6	5.5	5	4	3
1N485A	20	18	14	12	10	9	7	6
1N538	27	26	24	23	20	17	13	11
1N540	55	49	45	38	30	26	19	16
1N645	-	10	8	6.5	6	5	4	4
1N914	4	3	2	2	0.5	0.5	0.5	0.5
1N1692	160	180	170	150	130	115	80	60
1N1763(RAY)	-	60	40	35	33	30	25	22
1N1763(RCA)	-	40	32	28	26	24	20	17
1N2069	-	27	24	20	18	17	13	11
1N2071	-	18	16	15	12	11	8	7
1N4001	34	28	19	10	7	6	4	3
1N4002	15	10	8	7	6	6	5	4
1N4003	20	18	15	12	11	10	8	7
1N4004	35	31	26	22	18	16	12	10
1N4007	25	22	18	15	14	13	10	8
1N4148	3.5	2.5	0.8	-	-	0.5	0.5	0.5
1N5415	-	100	80	70	60	50	38	32
2E4	40	35	29	23	-	16	11.5	9
SR3011	-	20	12	10	9	8	6	5

FIGURE 1



Diode capacitance test circuit. C_x is a 5-pF ceramic reference capacitor.

correlate the data. I chose a 5-pF ceramic capacitor as a reference. My purpose here was to validate my measurement technique. The test circuit is shown in **Figure 1**. I used a 0.01-μF mylar capacitor between one diode lead and the meter terminal to provide DC isolation between diode and meter. Two 100-k carbon resistors provided capacitance isolation and DC from the power supply. Higher resistance values had no noticeable effect on the measurement. I assumed that, being reverse biased and drawing little or no current, the diode would have the same voltage across its terminals as the voltage measured at the power supply. All measurements were made below 1 MHz.

Observations

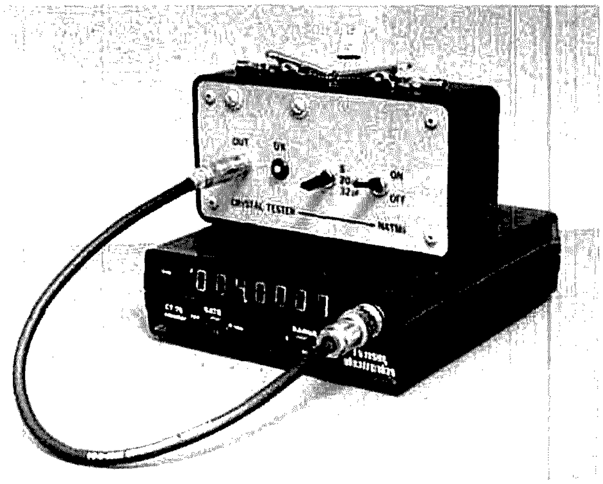
For VHF and UHF applications, it's desirable to test the diode's performance at the application frequency. I have noted experimentally that some diodes have a dissipation factor high enough to prevent their use in UHF applications. Also, the figure of merit (Q) of rectifier diodes doesn't seem to be as high as commercial varactor devices.

Although I didn't test it as part of this experiment, I assume that the base-collector junction of a UHF transistor would exhibit suitable varactor characteristics for UHF applications.

The applied RF voltage from a resonant circuit should be only a small percentage of the varactor's DC control voltage. Large RF voltage swings could cause the diode to conduct. This would result in waveform distortion and Q loss. Therefore, when used in a resonant circuit, the RF voltage applied to the diode would have to be expressed as a negative ratio from that on the inductor. A series-connected capacitor with a value nearly equal to the varactor will provide an approximate voltage ratio of 2:1. Zener diodes operated below their zener voltage value should exhibit varactor characteristics.

I couldn't measure the junction capacitance of microwave diodes like the 1N82A. The junction appeared to be conducting as a result of the AC signal from the capacitance meter, whether or not reverse DC bias was applied.

PRECISION CRYSTAL FREQUENCY CHECKER



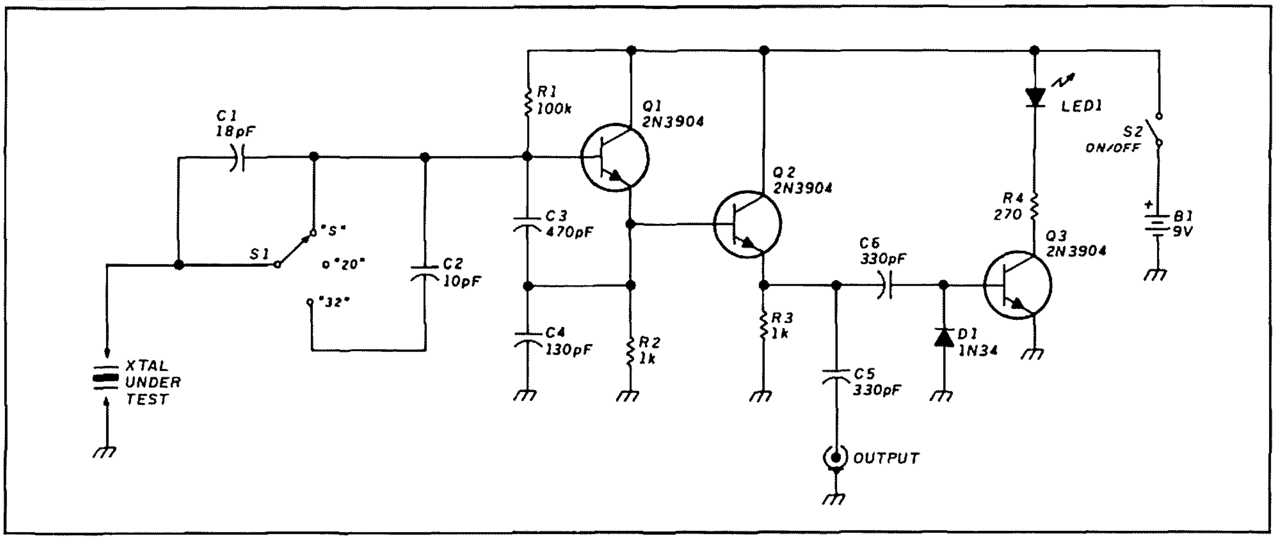
By Michael A. Covington, N4TMI, 285 Saint George Drive, Athens, Georgia 30606

Most crystal checkers perform a simple "yes/no" quality test or may give a relative indication of activity. I've designed one that teams up with a frequency counter to give precise frequency readings under three different load capacitances: series, 20 pF, and 32 pF. It can also measure inductance, though with a bit less accuracy. And even without the frequency counter, this checker will tell you whether or not a crystal oscillates.

This project really is a "weekender." You can find all the parts at Radio Shack, if they're not already in your junkbox.

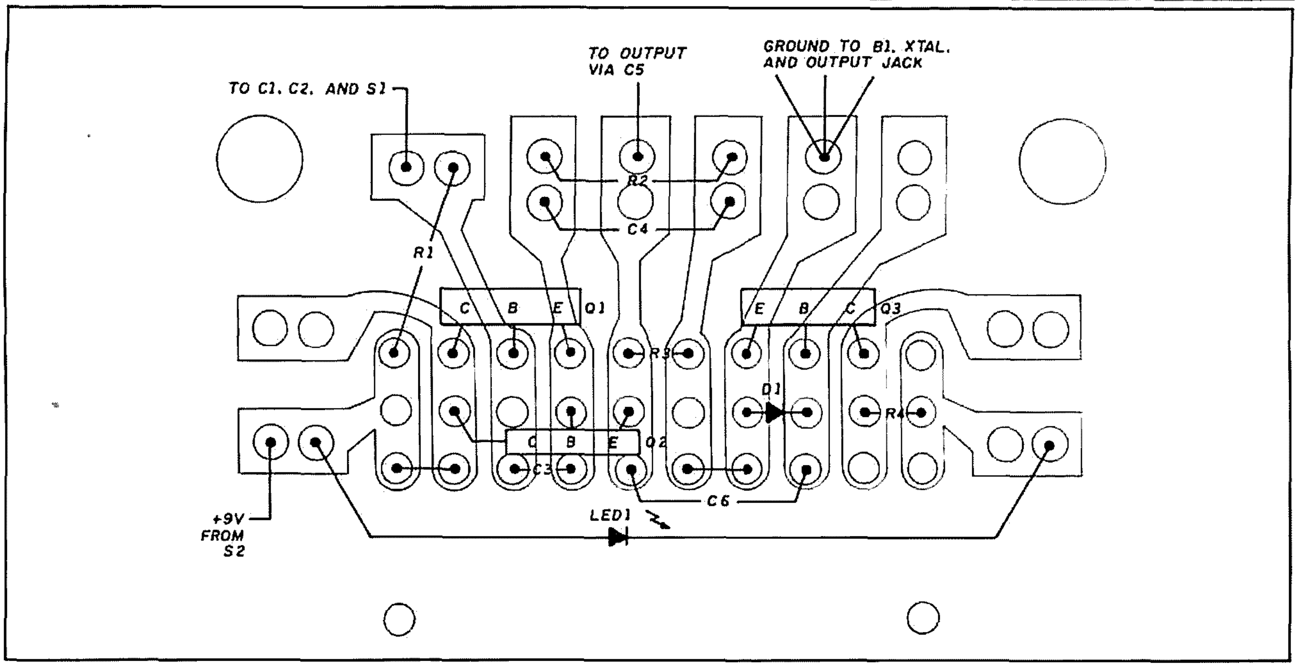
The circuit is a Colpitts oscillator (see **Figure 1**); capacitors

FIGURE 1



Circuit is a Colpitts oscillator with buffer amplifier.

FIGURE 2

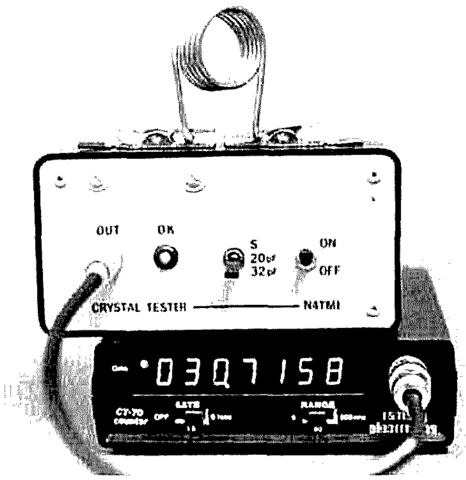


Parts placement diagram. View is from top (component side).

PHOTO A

Circuit board is half of a ready-made board from Radio Shack (RS no. 276-159A).

PHOTO B



To test a coil, measure resonant frequency with 32 pF.

were chosen to work with most crystals from 2 to 20 MHz. Q1 oscillates and Q2 buffers its output. For the yes/no test, D1 and Q3 rectify the signal from the oscillator and use it to light an LED. It's normal for this LED to dim or go out when there's a load (like a frequency counter) connected to the output jack.

When setting the load capacitance, the SPDT center-off switch S1 lets you connect the crystal directly to the oscillator, through an 18-pF capacitor, or through parallel 18 and 10-pF capacitors. Allowing for 2 to 4 pF of stray capacitance, this gives load capacitances near the nominal 20 and 32 pF. The direct connection gives a high load capacitance that puts the crystal very close to series resonance.

Most of the circuit is compactly built on half a Radio Shack 276-159A printed circuit board (see Figure 2 and Photo A). The tester is housed in a Bakelite™ box with a metal front. For reliable measurements, keep leads to S1 and the crystal as short as possible. Use alligator clips as a universal low capacitance crystal socket.

It's easy to test a crystal. Simply clip it in place, hook up the frequency counter, turn on the tester, and flip S1 to find out which load capacitance gives the correct frequency. This procedure also tells you how much the crystal can be "pulled" by changing the capacitance. The 20-pF load gives the highest frequency; the series connection gives the lowest. Overtone crystals will oscillate at the fundamental

TABLE 1

The frequency marked on a crystal isn't always the frequency at which it oscillates. The frequency measured by the crystal tester may be different yet, because the tester doesn't operate in the overtone mode. Here are some kinds of common crystals. F refers to the frequency marked on the crystal.

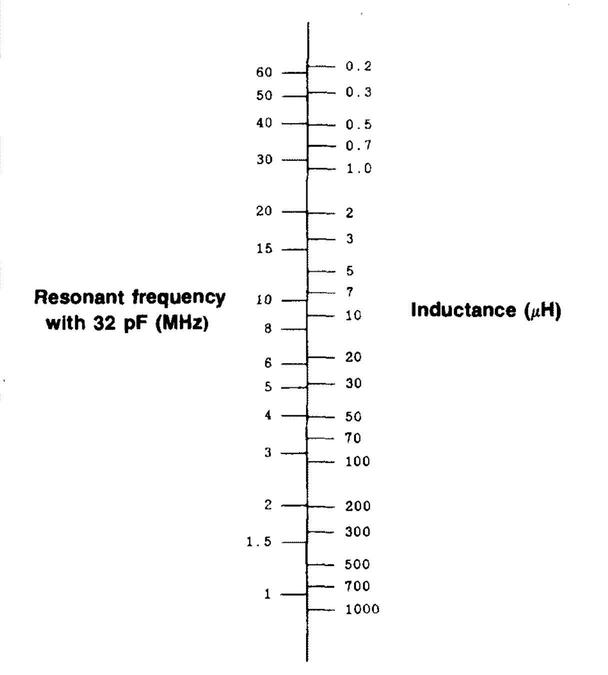
Type of crystal	Marked frequency (F, MHz)	Operating frequency (MHz)	Measured frequency (MHz)	Load capacitance
General purpose	1 to 20	F	F	Various
General purpose	20 to 60	F	F/3	Usually series
General purpose	55 to 100	F	F/5	Usually series
CB transmit	26.965 to 27.405	F	F/3	Series
CB receive*	26.510 to 26.950	F	F/3	Series
Scanner	30 to 50	F+10.7	(F+10.7)/3	Series
Scanner	140 to 175	(F−10.7)/3	(F−10.7)/9	Series
Scanner	440 to 470	(F−10.7)/9	(F−10.7)/27	Series
Scanner	470 to 500	(F−10.7)/10	(F−10.7)/30	Series

*CB receiving crystals are 455 kHz below the designated channel.

PARTS LIST

Capacitors (ceramic disk or polystyrene, 10 volts, ±20 percent)
C1 ***18 pF***
C2 ***10 pF***
C3 ***470 pF***
C4 ***100 to 150 pF***
C5,C6 ***330 or 470 pF***
Resistors (1/8 watt, ±5 or 10 percent)
R1 ***100,000 ohms***
R2,R3 ***1000 ohms***
R4 ***270 ohms***
Semiconductors
Q1,Q2,Q3 ***2N3904 (or equivalent) NPN silicon transistor***
D1 ***1N34 germanium diode***
LED1 ***Light-emitting diode***
Other
S1 ***SPDT center-off (three position) switch***
S2 ***SPST switch***
B1 ***9-volt battery***
Connector and holder for B1
RCA phono jack for output
Two alligator clips to hold crystal
Circuit board (Radio Shack 276-159A)
Enclosure

FIGURE 3



Computer-generated nomograph converts measured frequency to inductance.

frequency. For instance, a 27-MHz CB crystal oscillates at 9 MHz (see **Table 1**). Of course, you can use a crystal whose load capacitance is different from that for which it was ground or use an overtone crystal on its fundamental frequency.*

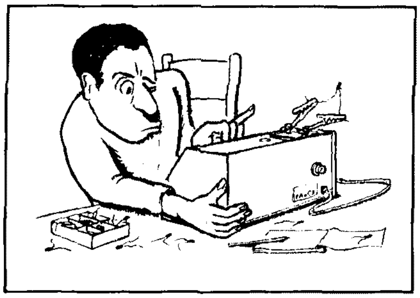
You can also test coils. Just connect a coil in place of the crystal, set S1 for a 32-pF load, and measure the frequency (see **Photo B**). Now find the inductance using the nomograph in **Figure 3**. You may find it more useful to remember the frequency than the inductance. It's the frequency at which the coil will always resonate with a 32-pF capacitor. Bear in mind that this measurement is inexact because the load isn't precisely 32 pF and the internal capacitance of the coil isn't taken into account.

*An overtone crystal oscillates near, but not exactly at, an odd multiple of the fundamental cut. Expect to see a small shift of several kHz when overtone crystals are operated in their fundamental mode. Ed.

Practically Speaking

By Joseph J. Carr, K4IPV

SAFETY STANDDOWN FOR THOSE WHO SERVICE HAM GEAR



I think it's a good idea to discuss matters of ham radio safety every so often, especially as they pertain to equipment and antennas. Unfortunately, from time to time we hear about a fellow Amateur Radio operator who was killed while working on a linear amplifier or installing an antenna.

A number of years ago, I wrote an article on electrical safety for *Ham Radio* (before I started writing "Practically Speaking"). A footnote added by the technical editor indicated that, during the week he was working on the manuscript, he needed to repair a high voltage DC power supply. The HV OUT line got loose, and snapped and barked around on the concrete floor like a venomous viper. (Hmmm...just where was that safety article?) If you've seen this material before, please bear with me. There are many who haven't read this information. Besides, a review of basic safety doesn't hurt anyone. (Even the United States Navy staged a one-day operations standdown for review of safety matters.)

The electronic equipment related to ham radio is inherently dangerous. If it's not used in accordance with some basic common sense rules, operation can lead to injury or even death. There are three situations to consider: *burns*, *macroshock*, and *microshock*.

Electrical accident incidents can cause first, second, and third-degree burns. These burns can occur in two ways: from the flash that results when an electrical arc occurs, or when current flows through body tissue. Any experienced emergency room physician can testify that ordinary 60-Hz power can cause burns. RF can also burn you. A physician, who also happens to be a ham, once told me that high power RF burns tend to be more serious because they penetrate deeper into the body. I recall one chap (wearing Bermuda shorts) who was working in his basement and sat on his 600-ohm parallel transmission line while he performed some adjustments. Unfortunately, he hadn't disconnected it from the transmitter. Someone else accidentally tripped the rig, intending to tune up and go on the air. After the cursing and screaming was over, the fellow found burns running parallel down the calf of his leg.

Macroshock is the type of electrical shock we all must take care to avoid. It results from direct contact with an electrical source. If you touch the 110-volt AC line while grounded, a very painful and possibly *fatal* shock will occur. Macroshock doesn't require the conduit of wounds or other breaches of the skin to enter the body.

Microshock is a more subtle form of electrical shock, and at one time it wasn't even recognized. However, the increased use of electrical equipment in hospitals during the fifties and sixties led some authorities to speculate that as many as 1,200 people a year were being accidentally electrocuted by tiny currents from hospital equipment that went unnoticed by the medical staff. Microshock is electrical shock caused by currents too minute to affect persons with intact skin, but able to inflict damage if introduced to the body through a wound. Microshock isn't normally a problem for hams.

For all forms of electrical shock to occur, there must be a difference in electrical potential between two points on the body. In other words, two points of contact must exist between the victim and the electrical source. That's why you sometimes see harmless "hair raising" exhibits, where people touch an electrostatic high voltage (>100,000 volts!) and their hair stands on end. These potentials are essentially "monopolar" with respect to the demonstrator, so no current flow exists. Similarly, some of the less prudent electricians will work a circuit "hot" (without turning off the power), feeling safe because they take care not to ground themselves or in any other way come between the hot wire and ground, or across two hot wires. Even so, *this is an extremely unsafe practice and must be discouraged at all times!*

In addition to electrical shock, there are other safety concerns you must consider when working with electricity. One major problem is fire. Overloaded or defective electrical circuits can spark, overheat, and/or cause a fire. Many fires every year are traced to faulty wiring or malfunctioning electrical equipment.

Electrical faults will also damage the equipment, the building where it is housed, or other equipment. A short circuit that isn't protected by a fuse may cause more damage in the shorted equipment, and may also affect building wiring and electrical components. In extreme cases, a fire may result. When fuses and circuit breakers aren't used, or are defeated ("penny in the fuse box" syndrome), there is a severe fire hazard and the level of damage done to any equipment involved will most certainly increase.

Less recognized, but nonetheless possible, is the hazard of *explosion* from electrical faults. There are at least two mechanisms which can cause explosions of this type. First, an overloaded circuit or electrical component may build up internal pressure (often from gas released when the device is severely overheated) and rupture. High power transformers and the main ripple filter capacitors inside high power RF amplifiers are types of equipment that can explode.

The second mechanism of explosion is sparking that occurs in the presence of flammable gases or vapors. If an electrical circuit is disconnected while operating, or if certain faults exist, then a spark may result. If that spark occurs when either flammable gases, oxygen, or vapors (like gasoline and certain waxes) are present, then a violent and dangerous explosion may result. Note

that oxygen isn't flammable itself, but vigorously promotes burning of other materials.

Besides the obvious danger of "shrapnel" wounds from the casing of an exploding device, there's also the possibility of injury from splattering boiling oil. In addition to the burns it can cause, this oil can be dangerous for other reasons. Certain older capacitors and transformers were built using PCB oil as an internal coolant. *PCB oil is a **potent** carcinogen.* Do not downplay the importance of this statement — PCB is dangerous stuff! Although most PCB-bearing electrical devices are now out of service, some are still around. Take care when handling older equipment. Be especially suspicious of elderly high power RF amplifiers. Should you find one of these devices, it would be a good idea to ask a competent person to dispose of the equipment. A PCB spill can close a building until a proper cleanup routine is completed; this can take a long time. Once when I published a PCB warning, a fellow wrote to me claiming the problem was overblown. I'll leave it to the experts; they still classify PCBs as dangerous.

What to do for the victim of an electrical shock

Death by electrical shock often occurs as a result of a phenomena called ventricular fibrillation (V.Fib.). This is an arrhythmic heartbeat; the heart merely quivers instead of beating. Unfortunately, a heart in V.Fib. is incapable of sustaining its blood-pumping effectiveness, so the victim dies within a few minutes — unless a person trained in cardiopulmonary resuscitation (CPR) is nearby.

Before *you aid a victim of electrical* shock, be sure that he is no longer in contact with the current, or that the current is turned off! Otherwise, when you touch him to administer aid, *you will also become a victim!*

As soon as the victim is clear of the electrical current, yell for help and initiate CPR. CPR won't bring him out of V.Fib., but it will provide life support until properly equipped and trained medical personnel can be summoned. They'll use a defibrillator to shock the victim's heart back into correct rhythm. They'll also use drugs and intravenous (IV) solutions to re-establish his body's balance.

None of these actions can be performed by an untrained person. In fact, even CPR can't be performed effectively by someone who hasn't learned the technique. Everyone who works near, on, or around electrical or electronic equipment should learn CPR. Teenage and adult family members should learn CPR, too. After all, who's going to save you if an electrical accident occurs at home in your hamshack or workshop? The local Red Cross, the Heart Association, some community colleges, and most local hospitals can direct you to certified CPR courses. It's impossible to learn CPR from watching medical shows on TV; get trained by a knowledgeable instructor!

How much current is fatal?

I once worked in a hospital electronics laboratory. One day I heard an intern claim that the 110 volts AC from a wall socket wasn't dangerous. Apparently he was told in medical school that it's not the voltage that kills; it's the current. I asked the doctor if he had ever heard of Ohm's law. According to Ohm's law, the current is the quotient of voltage and resistance, or $I = E/R$. It seems that doctor wasn't aware of this formula. The 110 volts AC available *in residential wall sockets is the most* common cause of electrocution in the United States. Also, medical studies reveal that the 50 to 60-Hz frequency used in AC power distribution almost worldwide is the most dangerous range of frequencies.

Higher and lower AC frequencies are less dangerous than 60-Hz AC, but they're not safe! Medical experts who've studied electrical shock say the killing factor is current density in a certain area of the right atrium of the heart called the sinoatrial node. Any flow of current through the body which causes a high level of current to flow in that section of the heart can induce fatal V.Fib. In general, the following rules of thumb are accepted for limb contact electrtical shocks through intact skin (macroshock):

1 to 5 mA	Level of perception
10 mA	Level of pain
100 mA	Severe muscular contraction
100 to 300 mA	Electrocution

Keep in mind that these figures are approximations and are *not to be accepted as guidelines to approximate "assumed risk."* Under certain circumstances, death can occur with considerably lower levels of current. For example, the risks escalate tremendously when you're sweating and standing in salt water.

Is high current at low voltage safe?

I once attended a design review meeting on a 100-watt commercial VHF mobile transceiver. One design specification called for insulation of low voltage (28 volts DC), high current (30 A) DC power supply terminals. One of the engineers present remarked that including this specification was like asking him to insulate the battery terminals of his car. His comment implied *that low voltage can never hurt you.* There are two false premises at work in his opinion.

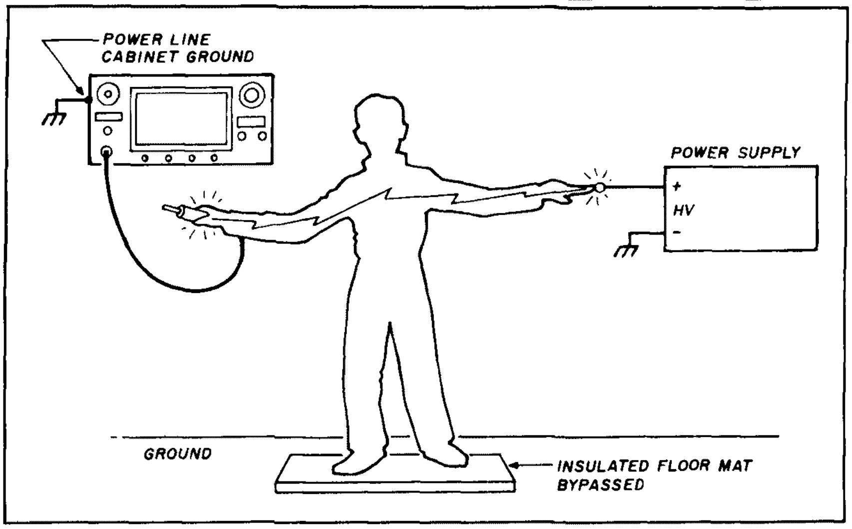
First, although low voltage, high current points rarely cause electrical shock, it's possible for a dangerous shock to occur when the person has a very low electrical skin resistance (very sweaty) or an open wound. I know of an electronics technician who was injured severely when he cut himself on a +5 volts DC, 30-A computer power supply terminal. Although this case didn't result in electroction, a large amount of current flowed in his arm, causing severe pain and some physical damage.

Second, high current is extremely dangerous if you are wearing jewelry! A two-way radio repair shop once used 12-volt batteries and battery chargers for the troubleshooting bench supply for mobile service. A technician working on the battery rack dropped a wrench, and it fell onto the battery making contact from (−) to (+) through his watchband. The large current turned the watchband red hot, giving him some serious second and third-degree burns. *Don't assume that low voltage, high current power supplies are harmless!*

Mechanisms of electrical shock

To raise your consciousness about how shock can occur, look at scenarios of electrical shock that might affect hams. **Figure 1** shows the direct aproach to fatal electrical shock. Imagine that you're grounded through conductive shoes and you touch an electrically hot point. You needn't be outdoors to be affected. A concrete garage, shop, or basement floor is a resonably good conductor, as are wet

FIGURE 1



Shock scenario where a grounded instrument probe creates a path for current.

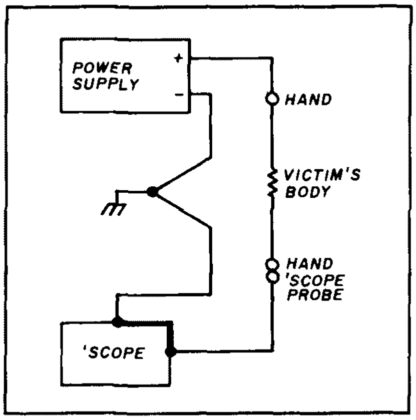
leather and some types of rubber shoes.

Figure 2 shows an indirect shock scenario that electronics workers should always keep in mind. Consider the grounded instrument probe (in this case an oscilloscope). When you grasp that probe, you may be grounded through the scope shield and the power cord ground conductor. If you touch a "hot" point, you'll get shocked — and may be killed.

A related scenario is shown in **Figure 3**. Here you see an AC/DC consumer appliance, like a low cost radio or TV set. Note that the oscilloscope probe ground is connected to the set ground, *which also happens to be one side of the AC power line*. Everything is fine as long as the AC plug is oriented correctly in the wall, and if the wall socket is wired correctly. But if you put the plug into the wall receptacle backwards, there will be an explosive short circuit which could electrocute the operator.

The fatal antenna erection job has contributed to the deaths of many hams. It isn't good practice to erect an antenna near a power line! NEVER! Every year we hear stories of people who were electrocuted when an antenna they were working on fell across the power lines, when they tried to toss a wire antenna over the power line in order to raise the antenna above them, or when a ladder they were using fell across the power lines. These foolish tactics will kill you. Incidentally, this is why OSHA-approved industrial

FIGURE 2



Equivalent circuit for *Figure 1*.

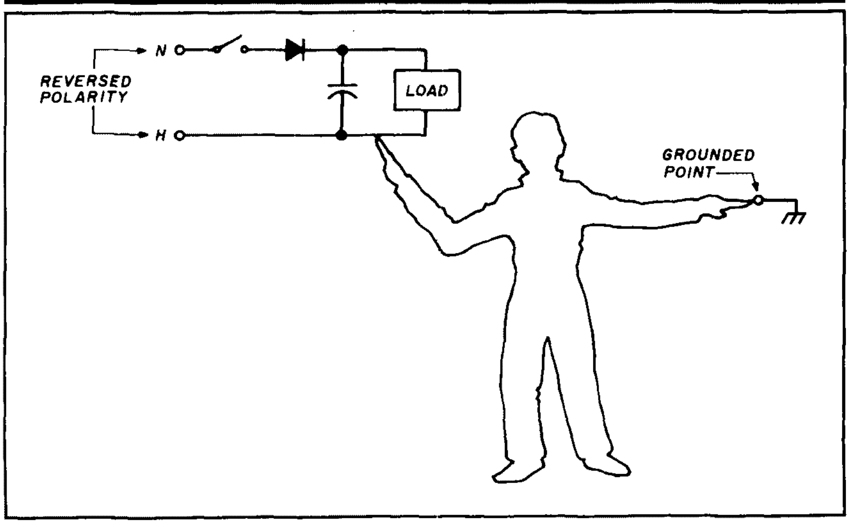
ladders are made of wood or other nonconductive material — not of aluminum like consumer ladders.

Some cures for these problems

Figure 4 is a schematic of the usual United States residential AC electrical system. Industrial electrical systems are a bit different at the service entrance, but become much like those in **Figure 4** when the power is distributed throughout the building. The power company distributes energy through high voltage lines. When it arrives at a point a short distance from the customer, it is stepped down in a "pole pig" transformer to 220 volts AC center tapped. The center tap (C.T.) of the transformer secondary is grounded, and therein lies the root of the problem. The two ends of the 220-volts AC secondary are brought into the building as a pair of 110-volts AC hot lines. Tapping across the two lines produces a 220-volts AC outlet; tapping from the ground line (i.e., transformer C.T.) to either hot line produces a 110 volts-AC outlet.

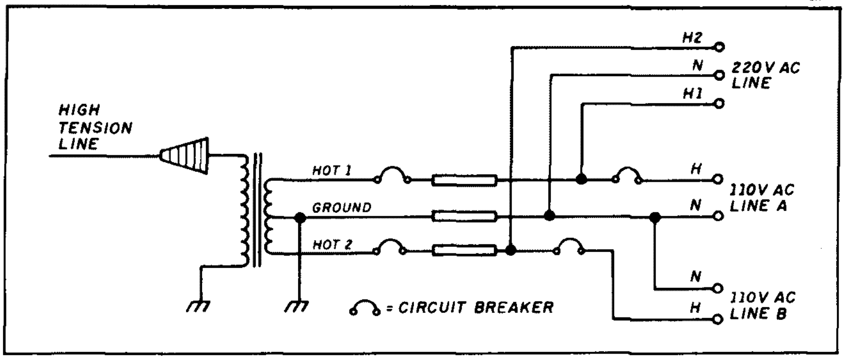
The electrical ground system in the United States is ground referenced; that's the problem. The solution is to make the little local electrical system non-ground referenced. This is done in hospital operating rooms, and in some intensive care units, for patient safety reasons. It should also be implemented on radio service benches, especially if AC/DC power supplies (damnable devices!) are ser-

FIGURE 3



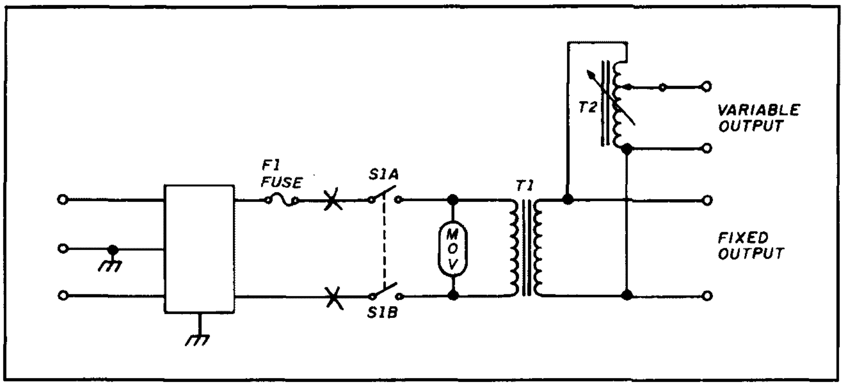
Shock scenario for AC/DC power supplies.

FIGURE 4



Residential AC electrical system.

FIGURE 5



Workbench electrical system.

viced. **Figure 5** shows the wiring for such a system. Transformer T1 is one of two types of isolation transformer. A 1:1 transformer gives a 110-volts AC isolated (non-ground referenced) AC line from a 110-volts AC standard line; a 2:1 transformer does the same thing from a 220-volts AC line.

The second transformer, T2, is an *autotransformer* used for varying the voltage on the AC line. It will typically allow you to set the output voltage from 90 to 140 volts AC with a 110-volts AC line voltage applied. This transformer is used by servicers to set the voltage higher or lower than normal to check radio operation or expose problems.

If you work on radio transmitters, you might want to place an electromagnetic interference (EMI) filter in the line at the points marked "X." The EMI filter is an LC-section that attenuates RF, but doesn't affect the 60-Hz power.

The "MOV" is a metal oxide varistor. It's used to clip the amplitude of high voltage line transients (100 microseconds or so) that could either damage or interfere with the operation of the equipment on the bench.

The circuit breaker or fuse protects the bench equipment and the transformer. It's always placed in the hot line, and can also be placed in both lines. However, fuses and circuit breakers should *never* be placed in the neutral line only. The switching shown in **Figure 5** breaks both lines. I prefer this approach because hot and neutral lines can be reversed accidentally, leaving you in the position of breaking a neutral while the hot line remains alive.

Some general points on safety

There's only one way to ensure that the AC line won't shock you — disconnect it. Make it your practice never to work on equipment that has the plug inserted into the power outlet. Don't trust switches, fuses, circuit breakers, or other people. If someone were to hand you a pistol, claiming that it was unloaded, the first thing you'd do is check it yourself. The same advice holds true for an electrical connection (which can kill you just as efficiently as a loaded and cocked pistol).

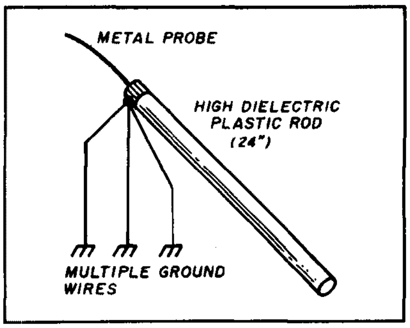
It's sometimes recommended that you work on high voltage devices with your left hand in your pants pocket. Supposedly, the "left hand to either leg path" is the most deadly. Even so, working with one hand in your pocket is awkward. I don't think anyone is able to work on a circuit safely with just one hand. It's better to use both hands, arrange a safe work environment, and use good techniques.

What's a safe work environment? It's one where the power system is isolated. The floor should be insulated by a carpet, treated masonite, a plastic cover, a rubber mat, wooden planking, or some other material. Also remember to keep the floor dry. Use an isolation transformer on the workbench for servicing radios.

When working on high voltage DC circuits, like those inside linear RF amplifiers, remember that *capacitors store electrical charge*. All filter capacitors must be discharged manually after the power is turned off. Also remember that *the capacitor must be discharged several times*. Even when a short circuit is placed across the capacitor terminals, all of the energy is not removed the first time it is discharged. Some energy is stored in the dielectric, even after the main charge is dissipated.

How do you short the capacitor terminals? With a screw driver? With an alligator clip lead? No! Use a shorting stick (**Figure 6**). *Do not make the stick out of wood!* Use a dielectrically competent plastic material instead. Be sure to use more than one ground line. If one of the lines falls off while you're working, there will be one or more lines left to carry off the charge. *Be sure to wear eyeglasses or safety goggles* to

FIGURE 6



Capacitor discharge wand.

prevent eye damage from the flying sparks.

The power cord ground is the only ground available to the equipment chassis when you have it running on the bench. This is insufficient for safety. My workbench has a grounded 5/16-inch bolt on the back edge. When I work on high power RF equipment, I ground the chassis to that ground point with a heavy cable. Don't overlook this ground. When only the power line ground is used, high power RF "travels" to strange places and can cause problems. I recall a time in a hospital electronics lab when we serviced an electrosurgery generator. These devices are high power (500 watt) 1-MHz (or so) RF power oscillators. (Some of them were spark gap "transmitters" as late as the early 1980s!) I forgot the ground rule and kicked the "transmit" switch. Bright blue RF sparks flew from every inch of the metal trimmed Formica™ workbench. Although the sight was spectacular, it scared the dickens out of me. (Note to Chuck, my ham colleague, who was there: no snickering!)

Conclusion

There's *no such thing* as complete, failure free safety protection. This statement is especially true when you're dealing with high power electrical devices like RF power amplifiers. But proper recognition of the mechanisms of danger and proper management of the risks will ensure that the environment is as safe as possible.

THE SERIES SECTION **TRANSFORMER**

By Charles J. Michaels, W7XC, 13431 N. 24th Avenue, Phoenix, Arizona 85029

Here's the problem. You've just put a half-wave dipole for 40 meters at 52 feet, connected 70 feet of RG-213 (50 ohm) coax, and trimmed the antenna to resonance at 7.15 MHz — but the best VSWR you can get is 1.6:1. Although the calculated extra 0.05-dB loss (above the 0.38-dB calculated matched loss — see **Appendix A**) is negligible, you don't like the VSWR and your transceiver doesn't either

Check the input impedance curve of a half-wave dipole versus height (0.375 wavelength in this case) in almost any antenna text and you'll find that you should have about 85 ohms for a VSWR of 1.7:1 at the antenna, and 1.62:1 through 70 feet of RG-213 line (see **Appendix B**). What to do?

How about a balanced L net at the antenna? Hmm, two coils and a capacitor in a weather protective box. Weight, wind load?

How about a stub matching system? Gosh, that RG-213 is heavy enough now!

Or how about a quarter-wave matching section? Let's see,

$$Z_o = \sqrt{Z_L\ Zi} = \sqrt{50 \times 85} = 65.2 \text{ ohms}$$

Sixty-five ohm coax? Not exactly a standard item!

Is a transmatch the only answer? No!

A solution

This, and many other matching problems, can often be solved by using a series section transformer. The technique uses a calculated length of feedline, ℓ_2, removed at a calculated distance, ℓ_1, from the load, and replaced by a piece of feedline with an impedance different from that of the main line (see **Figure 1**). This technique can also provide a match to loads that include reactance. Furthermore, it becomes *part* of the feedline rather than an addition.

Calculation

Use the aforementioned problem as an example. Your first task is to determine feasibility. The ℓ_2 section of line must have a characteristic impedance either less than

$$Z_1 / \sqrt{VSWR}$$

or greater than

$$Z_1 \sqrt{VSWR}$$

In this case the antenna VSWR is 1.7:1 and, because

$$50 / \sqrt{1.7} = 38.3 \text{ ohms}$$

and

$$50 \sqrt{1.7} = 65.2 \text{ ohms}$$

the ℓ_2 section must have an impedance of either less than 38.3 ohms or greater than 65.2 ohms. RG-11 at 75 ohms is standard, so the system is feasible.

To simplify the calculations, normalize the load impedance R_L and X_L, and the Z_2 impedance to the main line impedance Z_1 as follows:

$$n = Z_2/Z_1 \qquad \textbf{(1)}$$

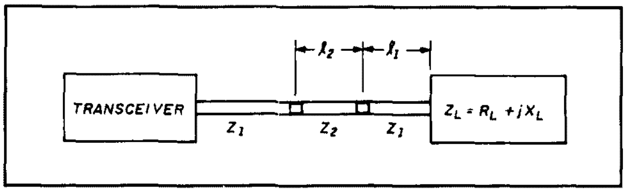
$$r = R_L/Z_1 \qquad \textbf{(2)}$$

$$x = X_L/Z_1 \qquad \textbf{(3)}$$

Because Z_2 is 75 ohms, R_L is 85 ohms, X_L is zero ohms (a dipole is a pure resistance at resonance), and Z_1 is 50 ohms using **Equations** 1, 2, and 3, n = 75/50 = 1.5, r = 85/50 = 1.7 and x = 0/50 = 0.

The *angular* length of section ℓ_2, or $\ell_2°$, is calculated as follows:

FIGURE 1



Series section transformer. Z_1 is the main-line characteristic impedance. Z_2 is the matching-section characteristic impedance. See text for calculation of Z_2, ℓ_1, and ℓ_2.

$$\ell_2{}^\circ = \arctan B \tag{4}$$

where:

$$B = \left[\frac{(r-1)^2 + x^2}{r\,(n - \frac{1}{n})^2 - (r-1)^2 - x^2} \right]^{1/2}$$

For the example:

$$B = \left[\frac{(1.7 - 1)^2 + 0^2}{1.7\,(1.5 - \frac{1}{1.5})^2 - (1.7-1)^2 - 0^2} \right]^{1/2}$$

$$= 0.842$$

A trigonometry table of tangents or a scientific calculator indicates that a tangent of 0.842 corresponds to an angle of 40.11°. This angle is converted to feet of transmission line by **Equation 5**.

$$\ell' = (2.733 \times \ell^\circ \times V_f)/F \tag{5}$$

where:

ℓ' = length in feet
ℓ° = length in degrees
V_f = velocity factor of line
F = frequency in MHz

In the example, velocity factor V_f is assumed to be 0.66 for both the RG-213 and RG-11. Quality coax is usually very close to specification. Using **Equation 5**:

$$\ell'_2 = (2.733 \times 40.11 \times 0.66)/7.15 = 10.12', \text{ or } 10'\ 1.4''$$

The angular length of section ℓ_1 is calculated as follows:

$$\ell_1{}^\circ = \arctan A \tag{6}$$

where:

$$A = \frac{(n - \frac{r}{n})\,B + x}{r + xnB - 1}$$

For the example,

$$A = \frac{(1.7 - \frac{1.7}{1.5})\,0.842 + 0}{1.5 + 0 - 1} = 0.441$$

0.441 is the tangent of 23.8°.

Again using **Equation 5**,

$$\ell'_1 = (2.733 \times 23.8 \times 0.66)/7.15 = 6', \text{ or } 6'0''$$

The design is now complete.

If you find the quotient is negative when calculating B of **Equation 4**, then Z_2 is too close to Z_1. This may happen when reactance is present in the load, despite initial indications of feasibility.

When calculating A of **Equation 6**, the result can be a negative number implying a negative angular length. In this case, add 180° to the negative angle to obtain the correct length.

Implementation

Cut the RG-213 coax line 6′ from the antenna and insert a 10′ 1.4″ length of RG-11 line. You can do this neatly with PL-259 connectors and barrels weatherproofed by wrapping the connectors and barrels with RTV compound. If you use connectors and barrels, include their lengths in the ℓ_1 and ℓ_2 lengths.

You may shorten the remaining RG-213 line to the station if you wish, as 10′ 1.4″ has been added. Because the addition is small, it's probably not worthwhile to shorten this line. In VHF applications, take extra care to include the connector and barrel lengths in the calculations.

Other applications

This system is applicable to both coaxial and balanced lines. In fact, because a wide range of balanced line impedance is available through your choice of conductor diameter and spacing, balanced lines offer a wide range of matching section parameters.

You can also use this system at the sending end, when it may be necessary to match a line of other than the 50 ohms for which your transceiver, VSWR meter, and low pass filter are designed.

Appendix A—Loss total

RG-213 at 7.15 MHz has 0.55-dB loss per 100 feet. Therefore, 70 feet of matched line has 0.7 × 0.55 = 0.38-dB loss.

$$\text{Loss total} = 10 \log_{10} \left[\frac{B^2 - C^2}{B\,(1 - C^2)} \right]$$

where

$B = 10^{Lm/10}$
Lm = loss matched in dB

and

$$C = \frac{S_\ell - 1}{S_\ell + 1}$$

where

S_ℓ = VSWR at load

$$C = \frac{1.7 - 1}{1.7 + 1} = 0.25926$$

$$B = 10^{0.38/10} = 1.09144$$

$B = 10^{0.38/10} = 1.09144$

$$\text{Loss/Total} = 10 \log_{10}$$

$$\left[\frac{1.09144^2 - 0.25926^2}{1.09144\,(1 - 0.25926^2)} \right] = 0.43$$

0.43 − 0.38 = 0.05 additional loss due to 1.7:1 VSWR

Appendix B—VSWR

S_i = VSWR at generator end of line
S_ℓ = VSWR at load end of line

$$C = \frac{S_\ell - 1}{S_\ell + 1}$$

$B = 10^{Lm/10}$
Lm = loss matched in dB

$$S_i = \frac{B + C}{B - C}$$

$$C = \frac{1.7 - 1}{1.7 + 1} = 0.25926$$

$B = 10^{0.38/10} = 1.09144$

$$S_i = \frac{1.09144 + 0.25926}{1.09144 - 0.25926} = 1.62$$

VSWR at input to line is 1.62:1

Ham Notebook

Cleaning Electronic Hardware

Try Tarn-X™ to clean and brighten tarnished or dirty coaxial connectors and similar electronic hardware. Tarn-X is claimed to contain acidified thiourea, detergent, and corrosion inhibitors. You can find it in department stores.

The manufacturer recommends wiping Tarn-X on with a cloth or cotton ball, but I've found it effective as a reusable dip. Many parts require dipping in Tarn-X for only 30 seconds to a minute and then rinsing in hot water. The rest can usually be cleaned up with a few swipes with an old toothbrush or fingernail brush and redipping.

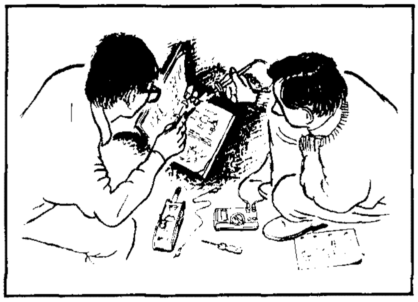
I've used this stuff for about 10 years, and my first bottle (I have two) is still going strong!

David McLanahan, WA1FHB

More on Elevated Radials: H.H. Beverage's 1921 Counterpoise System

Bill Orr's November 1989 column in *Ham Radio* touched upon the advantages of using elevated radials. This reminded me of an earlier article by H.H. Beverage, 2BML, that appeared in a 1921 issue of an RCA catalog. It described an aerial and counterpoise system, suggested by a Mr. Alexanderson, using a coupled ground wire. Details of 2BML's flat top antenna system of 1921 are shown in **Figure** 1.

Adding this coupled ground wire was the secret to improving antenna efficiency. Most Amateurs, and many broadcast stations (KGO was one of them) were already using the counterpoise at this time. Old boilers, model T cars, and other scrap metal buried 6 feet down served as the only antenna ground references for many stations until the elevated counterpoise became popular. Besides finding they were "getting out" better, operators also noticed that their antenna currents



were higher when using the counterpoise.

Beverage's system

Having permission to operate above 200 meters, 2BML chose to tune his flat top on 280 meters (1071 kHz). With a fair ground, his measurements showed a system resistance of around 70 ohms and 0.5-A antenna current. But with the elevated counterpoise attached, the system resistance dropped to 10 ohms!

The ground lead tap on the inductor was adjusted to cancel the capacitive reactance of the elevated counterpoise. The inductor was wound, using 3/8-inch tubing, into a 20-turn 15-inch diameter coil. Each time an adjustment was made the system was retuned for input power and frequency. With both the earth and counterpoise connected, a system resistance of 4 ohms and an antenna current of 8 A was obtained!

With 8 A of RF current going into the antenna, the counterpoise wire current (Ic) was about 6 A, and the earth ground current (Ig) was about 2 A. The counterpoise capacitance to ground was about 700 pF, and the antenna capacitance to ground was about 500 pF.

Conclusions

In the "old days" your antenna ammeter was your power meter. You put up an antenna, connected a ground and tuned for maximum antenna current. If you were "in the chips," you were the proud owner of a thermocoupled RF ammeter. You proudly told your listener, "I'm radiating 1.2 TCA amps* to my four-wire flat top up 80 feet, OM." This meant you were in between the Ford coil (spark) group, and the guys with a kW or two. When you decided to change from an "earth warmer" ground to a counterpoise, you became listed in the "Calls Heard" columns printed by *QST* every month. For March 1922 they listed six pages of logged heard or worked stations — spark or CW. Big DX was here!

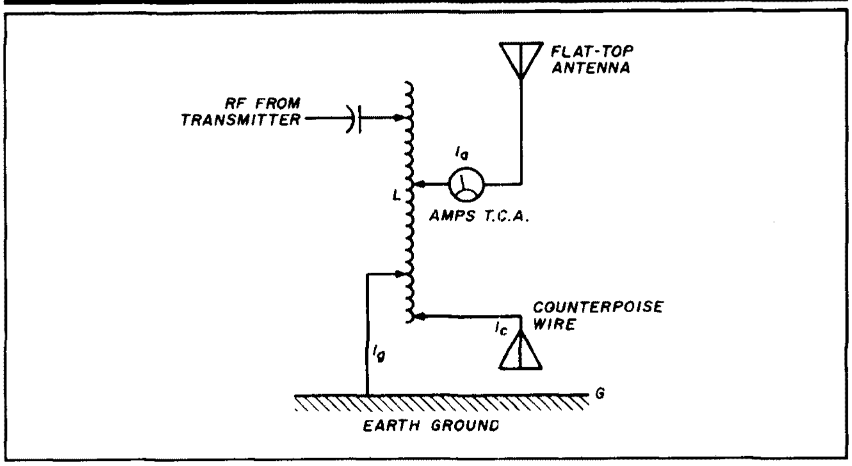
Going from a good ground to a counterpoise raised your signal about 10 dB. Adding the 2BML circuit gave you 3 or so *more* dB.

Take your pick. Rent a backhoe and bury a few thousand feet of copper, or put up a few elevated radials on 160 meters.

Dave Atkins, W6VX

*Amps measured by a thermo-couple ammeter. Ed.

FIGURE 1



2BML "flat-top" antenna system of 1921. Large inductor allowed matching anenna to transmitter and also permitted tuning-out capacitive reactance of counterpoise radial.

ANTENNA TUNER WITH A NEW TWIST

Bob Baird, W7CSD, 3740 Summers Lane, Klamath Falls, Oregon 97603

I find that the circuit commonly used in commercially built antenna tuners or transmatches leaves something to be desired (see **Figure 1**). Most use a tapped coil and selector switch. This can create problems, because on 10 meters a fraction of a turn on the coil may be critical. Also, you can't tap much closer than one turn. As compensation, the better units use a roller coil for continuous tuning. However, roller coils are expensive.

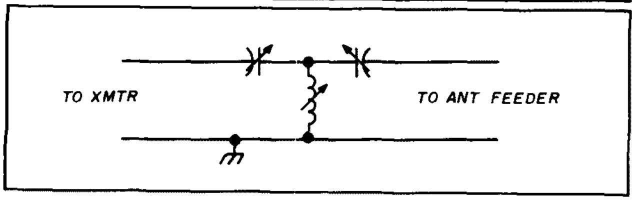
There's another problem with this circuit. It's basically a high pass filter. As Doug DeMaw and Bob Shriner have pointed out,[1] this sort of matching circuit does nothing for transmitter harmonics! I made note of a similar problem in an L network design some time ago. Older broadcast transmitters used "inductive" neutralization by means of a coil that was a bit too large in series with a large variable capacitor to achieve an effective variable inductor. Roller coils that operate at 10,000-volts RF or more are hard to come by. The large capacitor gave a small value of X_C and consequently a small value of IX_C, the voltage across the capacitor. This meant that the variable element for the neutralization process was a reasonably spaced capacitor which gave continuous tuning over a limited range (see **Figure 2**).

If you look at **Figure 3**, you'll notice that the series elements are inductive and the shunt element is capacitive. This is a basic *low* pass filter which will indeed do something for any harmonics present. Broadcast stations have been using this kind of matching network for years, for exactly the same reason. But you need two roller coils isolated from ground to build this circuit, and this would be expensive for the average ham.

How to do it inexpensively

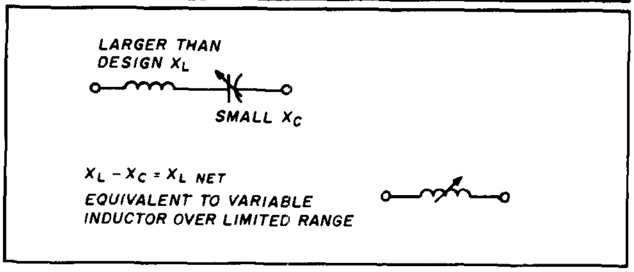
Suppose you were to replace the top of the "T" roller coils with tapped coils and series capacitors, giving a net variable L. The result is shown in **Figure 4**. Some time ago, I saw a nice forty-turn 2-inch diameter coil for a bargain price. I bought it, even though I had no particular project in mind. It turned out to be an Air-Dux no. 375-7433-PI, but any similar coil will do. I center tapped the coil and used it as the top of the "T." I was uncertain about the coupling effect of both

FIGURE 1



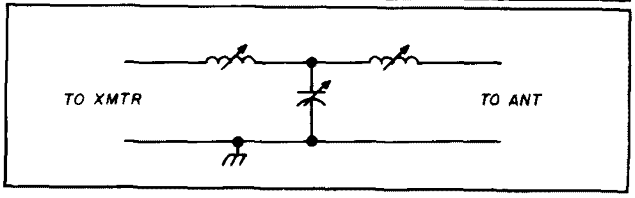
Popular and effective T match transmatch circuit used by many Amateurs. Although easy to build, this circuit is basically a high pass design *and offers no attenuation of transmitter harmonics.*

FIGURE 2



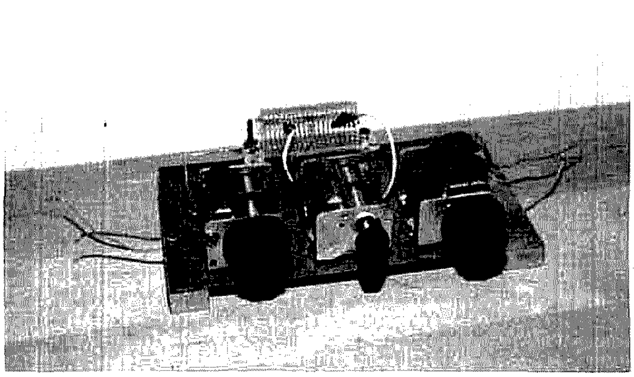
Making a fixed inductor variable. Capacitor in series with inductor is equal to smaller value inductor that is variable over a limited range. In practice a larger value inductor than the design calls for would be needed to compensate.

FIGURE 3



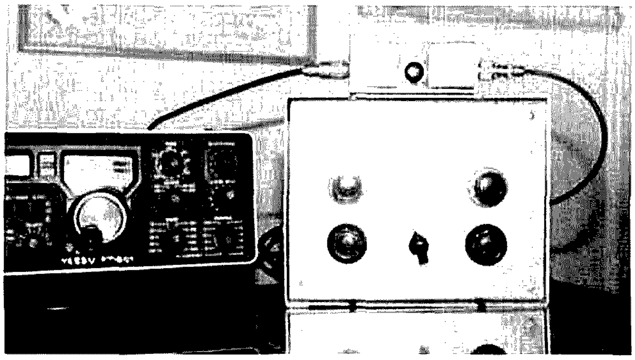
Low pass T match transmatch circuit offers good harmonic attenuation. Finding two suitable roller inductors can be both difficult and expensive.

PHOTO A



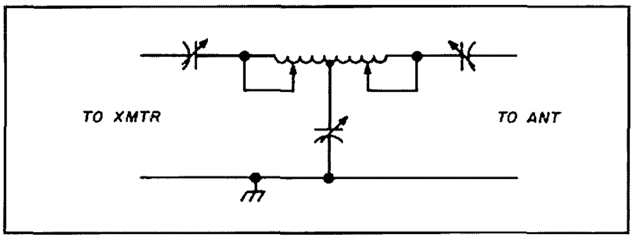
Author's breadboard version of tuner offers simple uncluttered construction and quick access to the traveling alligator clips. Note that insulated knobs and shaft extensions should be used on high power models to protect the operator from RF burns. W7IJK photo.

PHOTO B



Final version of tuner enclosed in shielded metal case offers better confinement of harmonic energy. Note that the capacitor bodies are RF hot and are mounted floating above ground. Cabinet is roomy enough to permit future installation of an internal SWR meter. W7IJK photo.

FIGURE 4



Roller inductor problem solved! Conventional Air-Dux coil, with alligator taps, is used instead of cumbersome roller inductors. The variable capacitors, salvaged from old broadcast receivers, allow finite adjustment of the inductor settings.

halves of the coil, but it worked fine. I used capacitors from old radios; many of them are available in lots of bargain sheets. Mine are two gang, but I only use one half. I used the breadboard in Photo A and traveling alligator clips in place of a rotary switch to determine if the circuit would really work. It did! A shielded box isn't absolutely necessary, but it might help reduce TVI. If your primary goal is to match a random length wire while camping, you can just use the breadboard.

I used the tuner briefly on an 80-meter dipole which had an SWR of 1.7:1 in the phone band, a 40-meter dipole, and a triband cubical quad with an SWR of greater than 1:1. I adjusted all of these easily to a 1:1 SWR. At my cabin in the woods I have a 400-foot wire, 60 feet high, fed on one end with open wire feeders (see *Ham Radio Horizons*, October 1979). I achieved a 1:1 match on all bands, 10 through 80, with this tuner.

Photo B shows the finished product in an aluminum box with two tap selector switches. The space at the top of the box is reserved for a built-in SWR meter. You needn't use a box this big; I simply had this one on hand.

High power

Obviously, this coil isn't meant for a kilowatt, nor are the capacitors. The unit works well with an FT 101. If you run a kilowatt of power, and are only interested in the 10, 15, and 20-meter bands, a ten-turn coil made out of 1/4-inch copper tubing will suffice. You could even use a hinged lid box and some husky traveling clips to avoid large contact selector switches. If you do this, you'll need medium spaced transmitting capacitors.

I hope you'll try this tuner. I've found that it really does discriminate against harmonics. hr

REFERENCES

1. Doug DeMaw, W1FB, and Bob Shriner,WA0UZO, "Matching the Transmitter to the Load," *QST*, February 1980, page 22.
2. Bob Baird, W7CSD, "Designing Impedance-matching Systems," *Ham Radio*, July 1973, page 58.

Tom McMullen, W1SL

Elementary Electronics — Wrapping Up AC Theory

In my April and May columns I looked at inductances and capacitors and the way they affect alternating current. This month I'll show you what happens when you have both of them in a circuit, and discuss some unusual terms that apply when working with complex circuits.

Current, voltage, and phase revisited

Remember that an inductance causes the voltage to lead the current through a circuit, while a capacitor has the opposite effect. But how do you calculate power in a circuit if the current isn't in phase with the voltage?

The term power factor, abbreviated pf (don't get this mixed up with picofarad, or pF) comes into play here. Power factor is usually expressed as a number less than 1, like 0.6, 0.85, and so on. Some people use it as a percentage like 55 or 92 percent. More about that later.

Power factor is determined by the ratio of true power to apparent power. **Figure 1** shows a circuit with an inductor, a resistor, and meters to measure applied voltage and the current that's flowing. Apparent power is determined by multiplying the voltage by the current, just as you do in a DC circuit. However, in this situation the result is expressed in volt-amperes, or VA, instead of watts. This difference is important, as you'll see in a moment. True power, as you might suspect, is what's being dissipated in the resistive part of the circuit.

Because the inductor stores energy in its magnetic field during one part of an AC cycle and releases it when the cycle reverses polarity, it doesn't dissipate any energy (unless the magnetic field intercepts something nearby). This means you can ignore the inductor as far as power consumed is concerned. You work only with the resistance. In the case shown in **Figure 1**, the ammeter shows 8 A flowing in the circuit; the resistor is 17 ohms. Using Ohm's Law, $P = I^2R$, $8^2 = 64 \times 17 = 1088$ watts. Note that you're using watts here, not volt-amperes. The power dissipated in the resistor is real power. You can feel it as heat when you touch the resistor. In circuits that don't have a physical resistor, there's still resistance present in wires in the inductance, transformer, electric motor, or whatever equipment you're using. The power dissipated as heat will show up as a warm (or overheated) inductor, motor, and so on. You can determine the power factor for this circuit. Apparent power equals E × I, or 200 × 8, or 1600 VA. You previously determined that the true power in the circuit is 1088 watts, so 1088/1600 = 0.673, or 67.3 percent.

So why bother with volt-amperes and power factor if you know the true power in the circuit? There are several reasons to be aware of VA and pf. It's very important to provide the correct size of wire and components in a circuit. Also, you don't want to overload the circuit's AC source (generator, amplifier, etc.). If, in the preceding example, you had planned for a wire size and generator size capable of handling true power (just over 1000 watts), the extra 600 watts could do real damage.

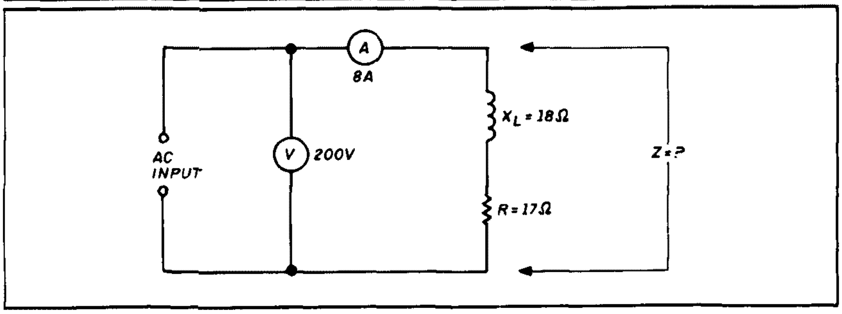
Here's another way to determine the power factor of that circuit. The circuit has an inductive reactance of 18 ohms and a resistance of 17 ohms in series. By using the information on vectors from my earlier columns, and doing a bit of intuitive reasoning, you might guess that the vector of these values would come out close to 45 degrees. (The vector is the hypotenuse of resistance plotted horizontally and the reactance plotted vertically. The resistance and reactance are only 1 ohm different in value). With a vector of 45 degrees, the impedance will be approximately 1.414 times either the resistance or the inductive reactance. Therefore, impedance (Z) = 1.414 × 18 = 25.45 ohms. You can work this out another way to see how close it comes. The formula says that:

$$Z = \sqrt{(R^2 + XL^2)} = \sqrt{(17^2 + 18^2)} = \sqrt{(289 + 324)} = 24.758 \text{ ohms.}$$

That's pretty close to our vector "guesstimate" of 25.45. Round this off to 25 ohms. With 200 volts applied to the circuit, current (I) = E/Z, or 8 A. This gives an apparent power of 200 × 8, or 1600 VA, confirming your first calculation.

You can find the true power with this information by obtaining the power factor. A different way to find the pf is to use the ratio of the resistance to the

FIGURE 1

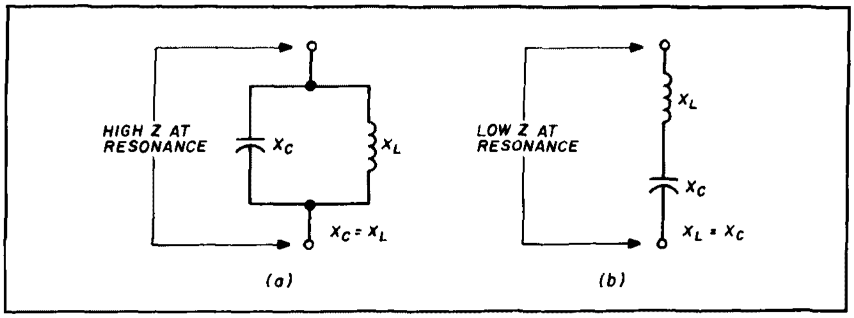


A simple AC circuit with inductance and resistance in series. The impedance, Z, is discussed in the text.

impedance:
pf = R/Z, or 17/25 = 0.68. Now find the true power by multiplying. VA × pf equals the true power in watts, or 1600 × 0.68 = 1088 watts — which is what the resistor appears to be dissipating. Using Ohm's law again, I = P/E = 1088/200 = 5.44 A appear to flow in the resistor.

Where's the extra current going? It's building up a magnetic field in the inductor in one half cycle, and returning to the circuit during the other half cycle. It's not lost, but does flow in the wiring and must be taken into account.

FIGURE 3



A resonant circuit results when the capacitive reactance and inductive reactance are equal but opposite. A parallel circuit is shown at (a); a series circuit at (b).

Inductance and capacitance combined

It should be apparent that if you can get the power factor back to near 100 percent, many of the problems will disappear. Everything I've just discussed in relation to inductive reactance applies to capacitive reactance. So if capacitors cause the current to lead the voltage, why not put a capacitor in the circuit to bring things back to normal? That's exactly what's done in many cases where a severe power factor problem exists.

I've placed a capacitor in series with the inductance in **Figure** 2. The capacitor has a reactance of 12 ohms. (Remembering the discussion of the "j" operator, you can say the reactance is −j12.) The impedance formula now becomes:

$$Z = \sqrt{(R^2 + XL - SC)^2}\text{ , or}$$

$$\sqrt{(17^2 + 18 - 12)^2}\text{, or}$$

$$\sqrt{(289 + 36)}, = \sqrt{325}\text{, or}$$

$$Z = 18.02\text{ ohms.}$$

Now, put Ohm's law to work on some other numbers:

I = E/Z = 200/18 = 11.1 A, and
apparent power = E × I = 200 × 11.1 = 2,220 VA, and
true power = I^2 × R = 123.2 X 17 = 2094 watts;
therefore,
power factor = 2094/2220 = 0.94, or about 94 percent.

That's a significant improvement. Note that the current flow has increased because the phase difference between voltage and current is smaller.

The resistance is now the more significant part of the circuit. By selecting the right value of capacitance, you can obtain a pf that's very close to 100 percent. As usual, circuit losses will prevent perfection.

That's heavy stuff and while not directly applicable to most of what you do in Amateur Radio, it's part of AC theory — a basic foundation for all electronics.

Resonant circuits

You've seen how inductances delay current flow and capacitances advance current flow. What happens when you connect them together, as in **Figure 3**?

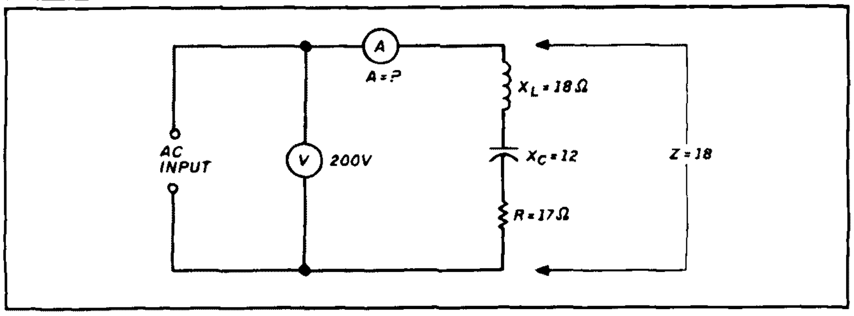
A simple explanation is that each makes up for the other's actions, or they balance each other out. Another explanation is that they form a resonant circuit. The requirement for resonance is that "the capacitive reactance and the inductive reactance must be equal and opposite at the frequency of interest." This works whether they are connected in parallel (**Figure 3a**) or in series (**Figure 3b**). This tuned circuit concept is basic to operation of radio equipment. It lets you separate one signal from another, reject a wide spectrum of signals while accepting others, and generate and amplify your transmitted signal while rejecting harmonic and spurious energy. The basic purpose of the circuit you need dictates which configuration you choose.

If you want a circuit that will reduce the signals outside a band while allowing those inside to pass, you want a parallel circuit (**Figure 3a**). This is because the parallel circuit appears as a high impedance to any signals at its resonant frequency. But off-resonance signals see a low resistance and are shunted to ground. By making either the inductance or capacitance variable, you can "tune" the circuit to a desired frequency, as is done in the front end of a receiver or in the output stage of a transmitter.

If you want to "trap" a frequency and keep it from passing, you can connect an L and C in series (see **Figure 3b**). A series resonant circuit has a very low impedance at its resonant frequency, and signals at this frequency will be shunted to ground while others will not.

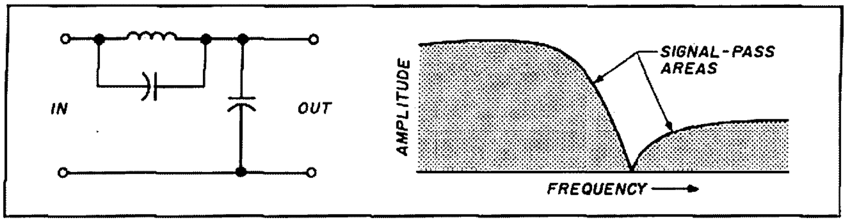
Most of the bandpass and band-reject filters used in modern Amateur

FIGURE 2



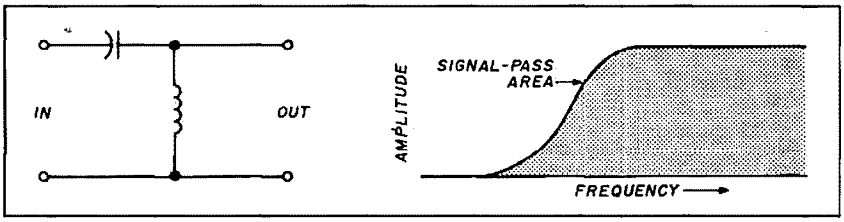
The circuit of *Figure 1* with a capacitor added in series. This changes the power factor and the current flowing, as shown in the text.

FIGURE 4



A circuit that allows low frequencies to pass while rejecting higher ones, commonly called a low pass filter.

FIGURE 5



A circuit that passes high frequencies while rejecting the lower ones, commonly called a high pass filter.

Radio equipment are variations of these circuits used in combination. These filters are used in the front end and IF stages to accept Amateur band frequencies and reject those out of the band. In the transmitter stages, they're used to get rid of harmonics and spurious signals before they reach the antenna. Modern low power filter circuits are all pretuned and switched in and out as you change bands. This means you must perform very few adjustments when changing frequency.

Figure 4 shows a typical low pass circuit and its resultant response; Figure 5 shows a high pass circuit and its response curve. By designing combinations of these circuits so that portions of the high and low pass areas overlap, you can create a bandpass circuit that allows only a selected part of the spectrum to pass while rejecting everything outside that band.

Further reading

Alternating current theory is far too involved to cover completely in these pages. If you choose to explore the subject further you'll encounter such interesting terms as conductance, admittance, and susceptance. You'll find ways to calculate the effects of parallel reactances, complex series reactances, and so on. For those of you who are interested, the various Amateur Handbooks have good chapters dealing with AC theory. *Radio Communications*, by R.L. Shrader (McGraw-Hill), is an excellent resource. Shrader includes self-exam questions designed to help those studying for either a commercial radiotelephone license or an Amateur radio license.

Coming soon: a packet radio update. Then I'll move on to other things you'll need to know to enjoy our great hobby. hr

DIVERSITY RECEPTION ON HF DIGITAL COMMUNICATION

Stephen M. Hall, WM6P, 664 Bristol Avenue, Simi Valley, California 93065

The use of diversity reception in radio communication isn't new to either commercial or Amateur stations. It's been used successfully since the 1940s to improve RTTY signal copy. In the past, this technique normally involved switching between two or more receivers based on strength of the received signals. In the reception of high frequency packet radio, signal strength alone is usually insufficient to judge the quality or completeness of a received packet of digital data.

I've conducted some experiments to evaluate the improvement in reception on HF packet radio when multiple antenna polarity diversity is used. Here are the results of those tests.

Initially, I assembled two parallel HF packet stations to measure and compare the performance of different modems and terminal node controllers (TNCs) available to Amateur packeteers. I extended these tests to include enhancements in reception using two antennas and two receivers, both tuned to the same frequency. The antennas I used in the test were a horizontally polarized four-element beam at 60 feet and a 1-wavelength Zepp, center fed with kW twin lead and a balanced tuner. The Zepp was supported by the common tower in a sloper configuration; polarization was primarily vertical. During my early tests, I used two Kantronics KAMs with individual video displays. I placed them side by side, so I could observe the performance of each parallel system in real time. The receivers I used in various parts of the tests included a Kenwood TS-930S, Kantronics KT-120, and Collins 51J-3 (R-388). I tested various modes, among them were AMTOR, RTTY, and packet.

I observed deep fades on both horizontal and vertical polarization, but the fade wasn't normally seen on opposing polarities at the same time. At times the fading was as great as 40 dB on 14 MHz; it would alternate or cycle between polarizations at rates of between approximately 1/2 to 4 seconds per cycle.

My observations of the two video terminals on RTTY and AMTOR modes revealed reception errors which alternated between displays. The incidence of simultaneous error reception was less frequent. It appeared that a large improvement in copy would be achieved if the two channels' signals could be merged into a single display. On packet, I saw frames from the same station alternating regularly between terminals. I was seeing unique packets — those copied on one TNC but not the other — on a single screen more frequently than I was seeing them on both terminals' screens simultaneously.

The use of an additional antenna which would take advantage of a second polarity appeared to be a technique worthy of greater investigation. After several hours of observation, I could see the obvious benefits of this technique. Fortunately, a program was developed which would let me measure the improvements using my computer.

This terminal program lets me gather statistics using an

IBM-AT computer. I can interface two communication ports on the computer to two terminal node controllers. The system can then count the total number of packets received by each KAM and the number of packets received uniquely by each one. I'm able to measure the improvement provided by an addition of a second channel or system.

I ran tests to measure systematic errors in the software, like the failure of the computer to properly count packets common to both channels and unique to each. No errors were observed. Then I ran further systematic tests to establish that all variations between channels would be due to antenna performance alone. I took off-the-air audio from a single receiver and fed it to both KAMs. I observed diversity between the KAMs alone; one unit consistently received 1 percent more packets than the other. I didn't find this surprising. What was surprising was that when two terminal node controllers were used in parallel, 14 percent more unique packets were received — even though the TNCs were demodulating the same audio. I was observing TNC diversity. These observations were consistent with those I had made using the two video terminals without the aid of the computer. The addition of a second TNC contributed 434 additional packets out of an average of 3064 received by the two units.

At this point, I connected the second antenna to allow polarity diversity reception. I ran tests at different times over several days on 14.109 MHz on the SKIP-NET packet network. I chose SKIP-NET because most of the stations in this net are generally in the pattern of both antennas. I pointed the beam northeast from my southern California receiving location. The Zepp antenna slopes in a similar direction, but the pattern is difficult to predict because of its wavelength.

The station received an average of 3344 packets using diversity; 2935 were received on the vertically polarized antenna and 3753 were received by the channel using the horizontally polarized beam. I believe that the difference in performance between the two antennas was due to antenna gain rather than polarization. I used both receivers and KAMs in a series of configurations to rule out the possibility that the differences were due to anything other than those contributed by the antennas. On the average, only 1305 packets of 3344 average total packets were received on both channels — an incidence level of 39 percent. Use of the second channel added 61 percent more unique packets. This seems to indicate that if the majority of net users were to incorporate this improvement, network loading would be significantly improved due to the elimination of many retries.

My observations on other digital modes appeared to demonstrate similar improvements, but I was able to gather statistics using fairly simple computer techniques only with packet. Because data with errors won't be presented to the output of the TNC, all data counted by the computer is known to be good. This isn't the case with monitored AMTOR or RTTY. Other techniques would have to be used with these modes to use a computer to gather statistics. If new terminal unit software were written for AMTOR which would allow users to take advantage of diversity, it's possible that even larger gains could be realized than those on packet, as individual AMTOR character groups could be compared or combined.

It would be impractical to use additional transmitter power or a single higher gain antenna to obtain improvements of this magnitude. Improving existing nondiversity performance by using more power would only load the frequency further and slow the passage of traffic on the network. An inherent advantage in using diversity is that it doesn't require that both stations participate. A single user may add the capability and enjoy the results independent of other stations. Other techniques that have been proposed to improve HF packet, like changes in baud rate, frequency shift, or modulation technique, would require substantial changes to all Amateur packet stations to maintain compatibility within the network. This would be difficult to achieve as there are large numbers of stations using the current standards.

Because there's more to communicating on packet than data reception, the TNC must acknowledge each received information frame while connected when this technique is used in other than monitor mode. If two TNCs are used, they must work in concert to maintain a proper connection with another station. This would require new software within the two interconnected TNCs, or a dual port TNC to monitor two incoming signals from two independent receiver/antenna systems. One TNC port would act as a primary; the second TNC would contribute packets not received on the primary channel. The TNC's microprocessor would compare data received on each channel and send acknowledgments for correct frames received on either channel.

Other statistics showed that, based on total packets received and independent of the use of diversity, the gain antenna received 27 percent more packets than the Zepp.

The experimental installation

I used the following hardware to add diversity to my station. When I included the second channel, I added a second receiver. I already owned a TS-930S, which has proved to be a good performer on HF packet. Kantronics provided a monoband KT-120 for my tests, and I also tested a Collins 51J-3 purchased at the Dayton Hamvention® for $100. Both proved quite suitable for packet reception. I used my existing 66-foot Zepp without modification with excellent results.

With suitable controller software, you need no other equipment. Only one video display is normally required for diversity packet or AMTOR, though I used two in the experimental configuration. Even when I grouped the lesser antenna, TNC, and receiver as the second channel, their contribution accounted for a 56-percent improvement when compared with the primary beam antenna, TNC, and better receiver.

Conclusions

Because I anticipated only modest gains using this technique, the tremendous improvement in reception which resulted was a surprise. I hope that suitable software for the currently available TNCs will be developed by the packet community or the manufacturers of multimode controllers which will allow diversity to be incorporated in HF packet stations. I have spoken with TNC manufacturers who wish to add this capability to their dual port designs. In some cases, this will be as simple as making firmware upgrades available on EPROM.

I wish to thank Phil Anderson, WØXI, and Karl Medcalf, WK5M, of Kantronics for their generous support of both time and equipment, which allowed me to pursue these experiments in digital diversity.

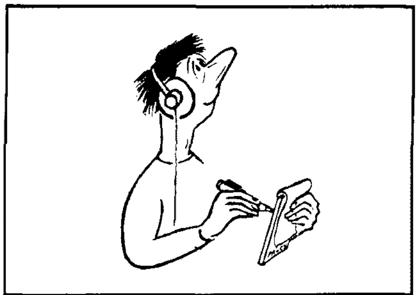
Garth Stonehocker, KØRYW

SUMMER SHORT SKIP SIGNALS

Each summer during June, July, and August, plus a week or so into May and September, short skip DX is enhanced. Short skip is propagation from the E region of the ionosphere, about 100 km above the earth. The geometry of propagation from a transmitter to a receiver, with a reflection height of 100 km tangent at the earth at each end (zero takeoff angle), defines the maximum distance for a hop at 2000 km (1200 miles). This enhancement results as the ionization produced near that height is moved horizontally (bunched together) by the electrojet current from the geomagnetic electric field into high density ion clouds. The clouds average in area about 40 by 500 miles wide and 6 miles thick.

These thin ion clouds are embedded in the regular E layer created during the daylight hours by the sun's ultraviolet light. The E layer supports a maximum signal frequency of 3.8 MHz at local noon during sunspot maximums. A signal can be increased up to 19 MHz and still be reflected if it hits a cloud because of the higher ion density and this reflection. Yes, reflection. The regular E is gradually bent down, or refracted, while the sharper edge of the thin cloud causes a more mirror-like reflection downward. The location of the ion clouds is similar to that of scattered weather clouds in the sky — hence the name sporadic E (Es). Because the E layer increases with the number of sunspots, the embedded Es clouds' ion density also increases. This is the only sunspot effect on mid-latitude Es. It is almost constant, and occurs every summer.

There are areas of the world where sunspots affect Es propagation because of increased geomagnetic disturbances. Larger numbers of higher intensity Es clouds develop where the geomagnetic and geographic equators are widely spaced in latitude — near Southeast Asia, South America, and Africa. The greater electrojet current that develops in these areas leads to this increased Es cloud



formation. Es also varies with the number of sunspots on the equatorward edge of the auroral zone, where particles enter the E region from the solar wind. The solar wind increases when the sun's spots flare (brighten). As a result, many clouds of Es which help with VHF auroral propagation develop in this region.

Now look at short skip Es signals. For a given distance, Es signal strength is an average of 12 dB greater than that of regular E signals as a result of the mirrored reflection. The signal strength is usually compared to the free space loss or signal decrease with distance at the frequency of transmission. A formula for relating this in decibels (dB) is:

$$L_b = 20 \log_{10} d + 20 \log_{10} f + 98.88 \qquad (1)$$

where d is distance in miles, f is frequency in MHz, and 98.88 is a constant set by a relationship to the reference antenna used in the measurements, the units used for d and f, and the normalized path attenuation in the locations where the measurements were made. Therefore, the formula can be used at any location, one hop distance, or any frequency band. Just subtract the dB for L_b from the free space loss of this formula:

$$L_f = 20 \log_{10} d + 20 \log_{10} f + 36.58 \qquad (2)$$

The units in **Equation 2** are the same as those in **Equation 1**. These formulas for the E region can be used to obtain the signal loss or, inversely, the signal strength's lowest value at a probability of near 10 percent.

Last minute forecast

The higher frequency bands will be best the first two weeks of June. The lower bands become strongest during the third and fourth weeks. Expect disturbances from the 5th to 8th due to solar flares, and from the 10th to 14th and the 20th to 24th as a result of decreasing solar flux and coronal holes. As the month passes, the build-up of thunderstorm noise will be increasingly evident towards the evening of each day. Es propagating modes will also build up during the month. The moon will be full on June 8th and at perigee (its closest approach) on June 21st. Summer solstice occurs on the 21st at 1533 UTC. The Aquarid meteor shower starts around the 18th, peaks about the 28th, and lasts until August 7th. The maximum radio echo rate will be 34 per hour.

Band-by-band summary

Six meters will provide occasional openings to South Africa and South America around noontime via multihop short skip E_s propagation.

Ten meters will have long skip conditions for many hours in the afternoon during the peak times of the 27-day solar cycle. Otherwise, look to sporadic E short skip and multihop openings around local noon for DX on this band. (Evening transequatorial openings usually don't occur in the summertime.)

Twelve, 15, and 17 meters, almost always open to some southern part of the world, will be the main daytime DX bands. Operate on 12 first; then move down to 15. DX is considered 5000 to 7000 miles on these bands. You may find some long one hop transequatorial propagation paths early in the month.

Twenty, 30, and 40 meters will support DX propagation from most areas of the world during the daytime and into the evening hours most days. DX on these bands may be either long skip to 2500 miles or short skip E_s to 1250 miles per hop. There are many good hours of DXing ahead during the long summer days.

Thirty, 40, 80, and 160 are all good for nighttime DX. Although the background thunderstorm noise is becoming noticeable, these bands are still quiet enough to provide good DX working conditions. Sporadic E propagation may contribute to enhanced conditions at local sunset and will occur more often during the next three months.

WESTERN USA

GMT	PDT	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	5:00	10	17	15	10	10	10	10	10
0100	6:00	10	17	15	10	12	10	10	10
0200	7:00	10	20	17	10	12	10*	10*	10
0300	8:00	10	20	17	12	15	10	10	10
0400	9:00	10	20	17	12	20	10	10	10
0500	10:00	12	17	15	15	20	10	10	10
0600	11:00	12	15	10	15	20	10	10	10
0700	12:00	15	15	10	15	20	10	10	12
0800	1:00	15	17	12	17	20	10	12	12
0900	2:00	17	20	20	17	20	12	15	15
1000	3:00	17	17	15	20	20	12	15	15
1100	4:00	20	15	12	17	20	15	17	17
1200	5:00	17	15	10	15	20	15	17	20
1300	6:00	15	12	10	12	20	17	20	17
1400	7:00	15	12	10	10	30*	17	17	17
1500	8:00	15	12	10	10	30	17	17	17
1600	9:00	15	10	10	10	30	17	17	20
1700	10:00	15	10	10	10	17	12	17	17
1800	11:00	17	10	10	10	15	10	20	17
1900	12:00	17	10	10	10	12	10	12	15
2000	1:00	20	12	12	10	12	10	10	15
2100	2:00	17	12	12	10	10	10	10	12
2200	3:00	15	15	15	10	10	10	10	12
2300	4:00	15	15	15	10	10	10	10	10

JUNE

MID USA

GMT	MDT	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN	CDT
0000	6:00	17	15	15	10	12	10	10	10	7:00
0100	7:00	17	17	15	10	15	10*	10*	10	8:00
0200	8:00	15	17	17	10	15	10	10	10	9:00
0300	9:00	15	20	17	12	17	10	10	10	10:00
0400	10:00	10	20	17	12	20	10	10	12	11:00
0500	11:00	10	30	15	15	20	10	10	12	12:00
0600	12:00	10	17	15	15	20	10	10	15	1:00
0700	1:00	10	20	20	15	20	10	10	15	2:00
0800	2:00	12	17	17	17	20	12	12	17	3:00
0900	3:00	12	17	15	17	20	12	15	17	4:00
1000	4:00	15	15	12	20	20	15	15	20	5:00
1100	5:00	15	15	10	17	20	15	15	17	6:00
1200	6:00	17	15	10	15	20	17	17	17	7:00
1300	7:00	17	12	10	12	30	15	17	15	8:00
1400	8:00	20	12	10	10	30	15	17	15	9:00
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1800	12:00	15	10	10	10	15	10	20	17	1:00
1900	1:00	15	12*	10	10	15	10	12	15	2:00
2000	2:00	15	12	12	10	15	10	10	15	3:00
2100	3:00	17	12	12	10	12	10	10	12	4:00
2200	4:00	17	15	15	10	12	10	10	12	5:00
2300	5:00	20	15	15	10	10	10	10	10	6:00

EASTERN USA

GMT	EDT	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE CARIBBEAN S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	8:00	12	15	15	10	15	10*	10*	10
0100	9:00	12*	15	15	10	15	10	10*	10
0200	10:00	15	17	17	10	17	10	10	10
0300	11:00	15	30	17	12	20	10	10	12
0400	12:00	15	20	17	12	20	10	10	15
0500	1:00	15	30	15	15	20	10	10	15
0600	2:00	15	20	15	15	20	10	10	17
0700	3:00	15	17	20	17	20	12	12	17
0800	4:00	15	17	15	17	20	12	15	20
0900	5:00	15	15	12	17	20	15	15	17
1000	6:00	12	15	10	15	20	17	17	17
1100	7:00	12	15	10	12	30	15	17	15
1200	8:00	10	12	10	10	30	15	17	12
1300	9:00	10	12	10	10	30	15	17	15
1400	10:00	10	12	10	10	30	17	17	15
1500	11:00	12	12	10	10	30	17	17	20
1600	12:00	12	12	10	10	20	15	20	17
1700	1:00	15	12*	10	10	17	12	20	17
1800	2:00	15	12	10	10	15	10	20	15
1900	3:00	17	12	10	10	15	10	10	15
2000	4:00	17	12	10	10	15	10	10	12
2100	5:00	30	12	12	10	15*	10	10	12
2200	6:00	17	12	15	10	15*	10*	10	12
2300	7:00	15	15	15	10	15	10*	10	10

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WANTED: Ham equipment and other property. The Radio Club of Junior High School 22 NYC, Inc. is a nonprofit organization, granted 501(C) (3) status by the IRS, incorporated with the goal of using the theme of Ham Radio to further and enhance the education of young people nationwide. Your property donation or financial support would be greatly appreciated and acknowledged with a receipt for your tax deductible contribution. Meet WB2JKJ and the "22 Crew" at Knoxville, TN Hamfest on June 2. Joe Fairclough will be the featured speaker telling the incredible story of the first 10 years of education thru communication at the core of the Big Apple. Write us at: PO Box 1052, New York, NY 10002. Round the clock hotline: VOICE (516) 674-4072, FAX (516) 674-9600.

BATTERY PACK REBUILDING: Don't pitch it—mail it—for FAST—PROFESSIONAL Rebuilding! Satisfaction guaranteed! ICOM: BP2/BP3 $19.95, BP5 $25.95, BP7/BP8/BP70 $32.95. KENWOOD: PB21 $15.95, PB21H $21.95, PB25/25H/26 $24.95. YAESU: FNB9 $19.95, FNB10 $23.95, FNB4/4A $36.95, TEN-TEC: $24.95 "U-DO-IT INSERTS" ICOM: BP3 $16.95, BP5 $22.95, BP7/8 $27.50, KENWOOD: PB21 $12.95, PB24/25/26 $19.95, TEMPO: S1-15/SERIES $22.95, YAESU: FNB4/4A $32.95, FNB10 $18.95, AZDEN: 300 $19.95, "NEW PACKS" ICOM: BP5 $43.95, BP8 (Base Charge Only) 39.95, YAESU: FNB2 $19.95, FNB11/FNB14 $44.95, SANTEC: 142/1200 $22.95. TELEPHONE/PAGER/COMPUTER/COMMERCIAL PACKS—Free catalog. Add $3.00 shipping/order. PA +6%. VISA/MC add $2.00. CUNARD, RD 6, Box 104, Bedford, PA 15522. (814) 623-7000.

VIBROPLEX COLLECTOR'S GUIDE. W1IMQ's new illustrated reference. Key Histories, identification guide, patents, more. 87 pages. Only $14.95 plus $2.00 s/h. Free info. Artifax Books, Box 88-H, Maynard, MA 01754.

KITS: QPR CW TRANSMITTER. 1 watt with crystal. 80m or 40m. $19.00 each. Code practice oscillator, $9.00. ($14.00 with key). Both great for Novice classes. SASE for info. WD8OYG (301) 586-2177. SR1, Box 2C, St. Leonard, MD 20685.

ATTENTION—HIRING! Government jobs—your area. $17,840-$69,485. Call 1-602-838-8885. Ext R15752.

TEN-TEC, New boxed latest 1990 production models, USA made 562 Omni V, 585 Paragon Transceiver, Titan 425 1.5 KW and 420 Hercules 1KW linear amplifiers, antenna tuners models 238, 253, 254, 239, 240 dry dummy loads, 2510B satellite station, mobile HF antennas, keyers, cabinets, filters and accessories. VISA/MC or check, for best deal write/phone Bill Slep (704) 524-7519, Slep Electronics Company, Highway 441, Otto, NC 28763.

1990 IDAHO HAM DIRECTORY. Calls, counties, phone numbers, names, addresses. $10.00. Ken Hastings, Box 434, Nampa, Idaho 83653.

WANTED: Rotary Inductor Johnson 226-1-4 or equivalent, 27.5 turns of 1/8" x 3/16" rectangular wire edge wound 3.5" OD. Spacing between turns increases toward one end. Old catalog listing 22.5 micro H. Calvin A. Hoerneman, 704 Apple Drive, Mechanicsburg, PA 17055. (717) 766-3465.

FOR SALE: Teletype Model 43 with books and one box of paper $100. New Wilson SY33 Tribander antenna and 33-6 MK 40-meter modification kit. $295. YOU SHIP. R. Davis, W9KK, Colfax, IN 46035. (317) 324-2442.

"SOFTSTART". Protect your valuable tubes and power supply diodes from DANGEROUS inrush surge currents. Can retrofitted into most amplifiers. Completely assembled Amateur net for "SOFTSTART" is $49.95 plus $5.00 shipping in the 48 states. OMEGA ELECTRONICS, 4209 Live Oak Road, Raleigh, NC 27604. (919) 231-7373.

FREE LIST of low cost ham equipment. Write to Jim Braddy, WA4DSO, 3037 Audrey Drive, Gastonia, NC 28054.

WANTED: BUY & SELL. All types of Electron Tubes. Call toll free 1-800-421-9397 or 1-612-429-9397. C & N Electronics, Harold Bramstedt, 6104 Egg Lake Road, Hugo, MN 55038.

CHASSIS AND CABINET KITS. SASE. K3IWK, 5220 Harmony Grove Rd, Dover, PA 17315.

ATTENTION: EASY WORK—Excellent Pay! Assemble products at home. Details. 1-602-838-8885. Ext. W15752.

RTTY JOURNAL published 10 times per year for those interested in digital communications. Read about RTTY, AMTOR, MSO'S, PACKET, RTTY DX and Contesting. Plus technical articles concerning the digital modes. $12.50 per year (foreign higher). RTTY JOURNAL, 9085 La Casita Ave, Fountain Valley, CA 92708.

CUSTOM MADE EMBROIDERED PATCHES. Your design, any shape and colors. Free ordering information, price list and random sample. HEIN, 79602 Manitou Trail, Glen Arbor, MI 49636-9711.

AVANTEK ATF10135, $12.00, MMIC's, P.C. board, SASE: WA3IAC, 7148 Montague St, Philadelphia, PA 19135.

"HAMLOG" COMPUTER PROGRAM. Full features, 18 modules. Autologs, 7-band WAS/DXCC. Apple, IBM, CP/M, KAYPRO, Tandy, C128 $24.95. HR-KA1AWH, POB 2015, Peabody, MA 01960.

IMRA International Mission Radio Association helps missionaries. Equipment loaned. Weekday net, 14.280 MHz, 1-3 PM Eastern. Nine hundred Amateurs in 40 countries. Rev. Thomas Sable, S.J., University of Scranton, Scranton, PA 18510.

WANTED: Henry VBC 3000. Dick (716) 366-4092.

INTERESTED IN PUBLIC SERVICE? Join your Local Radio Emergency Associated Communications Team. In Pennsylvania call (717) 938-6943 or write REACT, 1160 Old Trail Rd, Etters, PA 17319.

RUBBER STAMPS: 3 lines $5.00 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

ELECTRON TUBES: Receiving, transmitting, microwave... all types available. Large stock. Next day delivery, most cases. DAILY ELECTRONICS, PO Box 5029, Compton, CA 90224. (213) 774-1255.

ATTENTION: EARN MONEY READING BOOKS! $32,000/year income potential. Details 1-602-838-8885. Ext Bk15752.

RECONDITIONED TEST EQUIPMENT $1.25 for catalog. Walter, 2697 Nickel, San Pablo, CA 94806.

COMING EVENTS

Activities — "Places to go . . ."

SPECIAL REQUEST TO ALL AMATEUR RADIO PUBLICITY COORDINATORS: PLEASE INDICATE IN YOUR ANNOUNCEMENTS WHETHER OR NOT YOUR HAMFEST LOCATION, CLASSES, EXAMS, MEETINGS, FLEA MARKETS, ETC, ARE WHEELCHAIR ACCESSIBLE. THIS INFORMATION WOULD BE GREATLY APPRECIATED BY OUR BROTHER/SISTER HAMS WITH LIMITED PHYSICAL ABILITY.

JUNE 1-2: GEORGIA: 1990 ARRL Georgia State Convention and SE Packet Conference, Heritage House Motel & Convention Center, Albany. Admission $3. Free parking. For information Albany ARC, POB 1205, Albany, GA 31702. (912) 883-7910, M-F, 9-5.

JUNE 2: ONTARIO: Central Ontario Amateur Radio Flea Market, Bingeman Park, Kitchener, Ontario, Canada. Contact Ray Jennings, VE3CZE, 61 Ottawa Cres., Guelph, Ontario, N1E 2A8, Canada. (519) 822-8342.

JUNE 2: KANSAS: The Pittsburg Repeater Organization's annual Hamfest, Lincoln Center, 709 West 9th Street, Pittsburg. 8 AM to 3 PM. Admission $5.00 includes chicken dinner. Contact Ken Johnston, 2402 Wall Avenue, Joplin, MO 64804. (417) 623-1895.

JUNE 2-3: WASHINGTON: The Apple City ARC's Hamfeast, Rocky Reach Dam, Wenatchee. *Registration:* Amateurs $5; Others $1. Under 12 free. Free camping and trailer spaces, many area attractions. Mail registration to Bob Lathrop, Treasurer, 919 N. Woodward Drive, Wenatchee, WA 98861.

JUNE 3: NEW YORK: Hall of Science ARC's Electronic Hamfest and Computerfest, Hall of Science Parking Lot off 47th avenue and 111th Street, Flushing Meadow Park, Queens. 9 AM to 3 PM. Donation Buyers $3. Sellers $5/space. For information call Steve Greenbeaum, WB2KDG (718) 898-5599 or Phil Kubert, N2HYE (213) 777-8648 EVENINGS ONLY.

JUNE 3: CONNECTICUT: The Newington ARL's 7th annual Ham Radio and Computer Flea Market, Newington HS, Rt 173, Willard Avenue. 9 AM to 2 PM. Admission $3. For information SASE to NARL Hamfest, PO box 165, Pleasant Valley, CT 06063. (203) 523-0453.

JUNE 9: MICHIGAN: The Independent Repeater Association's annual Hamfestival, 44th Street Armory, Byron Center. 8 AM to 4 PM. Admission $3/advance; $4/door. For information write I.R.A., 562—92nd Street SW, Byron, Center, MI 49315.

JUNE 9: MICHIGAN: CMARA's 16th annual Hamfest, Midland Community Center, Midland. 8 AM to 1 PM. Admission $3. For information SASE to CMARA Hamfest, PO Box 67, Midland, MI 48840. (517) 631-9228 evenings and weekends.

JUNE 9: NORTH CAROLINA: The Forsyth ARC's 3rd annual Hamfest, Computer and Electronics Fair, Benton Convention Center, 301 West 5th Street, Winston-Salem. 9 AM to 3 PM. Admission $4/advance; $5/door. Wheelchair accessible. For information SASE to Jim Rodgers, N1DRI, W-S Hamfest, PO box 11361, Winston-Salem, NC 27116 or call (919) 760-2493, 9 AM to 10 PM.

JUNE 10: ILLINOIS: The Six Meter club of Chicago will hold its 33rd annual Hamfest, 91st and Wolf Road, Willow Springs, southwest of Chicago. Registration $3/advance; $4/gate. Gates open 6 AM. Advance tickets from Mike Corbett, K9ENZ, 606 South Fenton Avenue, Romeoville, IL 60441 or any club member.

JUNE 10: ILLINOIS: The Egyptian Radio Club's EGYPTIANFEST, at their clubgrounds, Chouteau Place Road, Granite City. 6 AM to 3 PM. Contact Carl Waller, WB9YDK, PO Box 562, Granite City, IL 62040. (618) 345-6469 for details.

JUNE 10: OHIO: The Goodyear ARC's 23rd annual Hamfest and Family Picnic, Wingfoot Lake Park, Akron. 8 AM to 4 PM. Family admission $4.00/advance; $5.00 at gate. For tickets and information contact William F. Dunn, W8IFM, 4730 Nottingham Lane, Stow, OH 44224. (216) 673-8502.

JUNE 10: PENNSYLVANIA: The 17th annual Central PA Ham and Computerfest, sponsored by the Milton and Central Susquehanna ARCs, Winfield Fireman's Fairgrounds. 8 AM to 5 PM EST. Donations $4/gate. For information Jerry Williamson, WA3SXQ, 10 Old Farm Lane, Milton, PA 17847. (717) 742-3027 or Bob Stahl, KQ3KR, 452 Fourth Street, Northumberland, PA 17857. (717) 473-7050.

JUNE 16: NEW JERSEY: The Raritan Valley Radio Club's 19th annual Hamfest and Computer Show, Watchung Hills HS, Warren. Gates open 8 AM. Lookers $4 donation. Spouse and kids free. For further information call Dave, KA2TSM (201) 763-4849 or John, WA2C (201) 968-5070.

JUNE 16-17: MONTANA: The Valley ARC of Glasgow is sponsoring a Father's Day Ham Picnic. Contact Linda Gray, N7FSH (406) 228-2324.

JUNE 17: MARYLAND: The Frederick ARC's annual Hamfest on Father's Day, Frederick County Fairgrounds. 8 AM to 4 PM. Admission $4. Spouse and kids free with one paid admission. For information Ernie Hansen, K3VVV, PO Box 589, Mt. Airy, MD 21771.

JUNE 17: INDIANA: The Lake County ARC's annual Fathers' Day Hamfest, Industrial Building of Lake County Fairgrounds, Crown Point. 8 AM to 2 PM. Admission $3.50. For information contact Ken Brown, KE9TC, 918 Chippewa, Crown Point, IN 46307. (219) 663-5035.

JUNE 17: MICHIGAN: The Monroe County Radio Communications Association's 1990 Hamfest, Monroe County Fairgrounds, Monroe. Tickets $3/advance; $4/gate. Handi parking and wheelchair accessible. Contact Fred Lux, WD8ITZ, PO Box 982, Monroe, MI 48161. (313) 243-1053.

JULY 4: PENNSYLVANIA: Firecracker Hamfest sponsored by the Harrisburg RAC, Bressler Picnic Grounds, Harrisburg. Starts 8 AM. Admission $3. Contact Dave Dormer, KC3MG (717) 939-4957 for reservations.

LAUREL ARC monthly (except December) Amateur exam sessions for all license classes. No fee is charged. Preregistration is required. Call (301) 725-1212, Maryland Radio Center, 8576 Laureldale Drive, Laurel, MD 20707.

THROUGHOUT 1990 the Major Armstrong Memorial Amateur Radio Club (MAMARC) will sponsor events commemorating Major Edwin Howard Armstrong's achievements in the field of radio broadcasting. The club is seeking other Amateur operators around the world who are willing to research Major Armstrong's accomplishments and become official MAMARC special events stations. Major Armstrong was a pioneer responsible for the creation of Wideband FM and the inventor of the superhetrodyne receiver. If you are interested in participating and becoming an official MAMARC special event station contact Barry Groupp, N2HDW, MAMARC, c/o *100th Birthday Committee, PO Box 581, Alpine, NJ 07620.* Please SASE.

YOUTH LINK NET. Open to all Hams under age 18. Saturdays at 2000 UTC, 28.425 MHz. For more information contact Net Control, George Manning, WB5NMH, 602 Glendale St, Burkburnett, TX 76354.

FREE 1989-90 *Florida* two meter repeater directories are currently being distributed by the Hernando County Amateur Radio Assn. of Brooksville, FL. Ask for one at any official Florida Welcome Center or SASE to Repeater Directory, Hernando County ARA, POB 1721, Brooksville, FL 34605-1721.

AMATEUR EXAMS. July 14, September 8, November 17. St. Mary Medical Center, 3333 No. Seminary Street, Galesburg, IL 61401. 12 Noon to 2 PM. For information contact Larry Heller, KA9PCU, 1436 Brown Avenue, Galesburg, IL 61401. (309) 342-5977.

Monthly Ham Exams. The MIT UHF Repeater Association and the MIT Radio Society offer monthly ham exams, all classes Novice to extra: next-to-last Wednesday of each month. (JUNE 20) 7:30 pm, MIT Room 1-150, 77 Mass Avenue, Cambridge, MA. reservation requested a couple of days in advance, walk-ins welcome; call the shack (617) 253-3776, or Nich Altenbernd (617) 437-0320. Exam fee $4.95. Bring copies of your current license (if any) and Certificates of Completion (if any), two forms of picture ID, and a completed form 610, available from the FCC: (617) 770-4023.

AMATEUR RADIO CLASSES: For those people interested in obtaining a Novice (basic level) Ham license or upgrading to Tech/General, the Chelsea Civil Defense, in cooperation with QRA Radio Club, will sponsor Amateur Radio Communications classes evenings at Chelsea High School starting MARCH 7, 1989. For more information write Frank Masucci, K1BPN, 136 Grove Street, Chelsea, MA 02150. Please enclose your telephone number.

Cushcraft D3W Rotatable WARC Dipole and Ten-3 10-Meter Yagi

Cushcraft's new D3W world ranger dipole covers 30, 17, and 12 meters on the new WARC bands. It's a sturdy, rotatable dipole that's easily mounted on any mast from 1-1/2 to 2 inches in diameter with your existing triband or other antennas. The D3W features high performance, high-Q traps, heavy wall tubing, and rugged stainless steel hardware.

This rotatable WARC dipole features automatic frequency selection for either 12, 17, or 30 meters, is rated for 2000 watts PEP, is 34 feet long, and weighs 11 pounds.

The Cushcraft Ten-3 is a high performance 8-dB three-element Yagi with 8-dB forward gain. The beam also offers a front-to-back ratio of 25 dB.

The Ten-3 has an 8-foot boom, and takes a mast size of 1-1/2 to 2 inches, and can be installed on a simple mount with only a light rotator. The reddi match system provides 50-ohm feed for a standard PL259 connector. The antenna is power rated for 2000 watts PEP. All tubing is heavy wall, hard drawn, bright finish aluminum.

For additional information contact Cushcraft Corporation, PO Box 4680, 48 Perimeter Road, Manchester, New Hampshire 03108. Telephone (603)627-7877.

Circle #301 on Reader Service Card.

Ameritron AL-82 Linear Amplifier

Ameritron announces its full legal power linear amplifier with two 3-500Z transmitting tubes. The AL-82 features dual illuminated meters. The grid current meter gives a constant reading; the multimeter displays plate voltage and current, peak RF output power, and drive power/ALC.

The Pi-L tank circuit permits full impedance matching over the 160-meter band. The tuning capacitors and bandswitch have a 35-percent safety factor to avoid tank circuit component failure.

The cooling system keeps the components and 3-500Z tubes safely below the manufacturer's ratings, even while operating at 1500-watts output with a steady carrier The filament supply has inrush current limiting to ensure maximum tube life. Complete shielding and bypassing helps prevent TVI and RFI.

The AL-82 covers 160, 80, 40, 20 and 15 meters and gives 80-percent rated output on 12 and 17 meters. It can be modified to cover 10 meters upon presentation of an Amateur license. An export model is also available.

For more information write Ameritron, 921 Louisville Road, Starkville, Mississippi 39759 or call Ameritron at (601)323-9715, FAX (601)323-6551.

Circle #302 on Reader Service Card.

ELNEC Advanced Antenna Analysis Program

Roy Lewallen, W7EL, announces ELNEC — a new full-featured antenna modeling and analysis program for PC-compatible computers. ELNEC eliminates the tedious and error-prone procedure of counting "pulses" to determine where sources and loads are placed that is found in other MININEC-based programs. It also provides true current sources for phased array analysis. The program was designed to be user friendly and is entirely menu driven. Users can add, delete or modify wires, sources, loads, and ground media with a few keystrokes while the program automatically keeps sources and loads where you originally placed them. Other features include azimuth and elevation plotting; saving and recalling antenna descriptions; beamwidth, gain, sidelobe, front/back and front/sidelobe analysis, current distribution, and SWR (50- or 75-ohm systems). Plots can be printed on Epson-compatible 8/9 or 24 pin dot-matrix printers. All features except plotting are available on nongraphics systems, and pattern data are presented in tabular form. Plotting requires CGA, EGA, Hercules, or compatible capability. Two versions are available, optimized for coprocessor or non-coprocessor systems (specify when ordering). Price is $49, postpaid (United States, Canada, Mexico) from Roy Lewallen, W7EL, PO Box 6658, Beaverton, Oregon 97007.

Circle #303 on Reader Service Card.

SA Series Programs for IBM PCs

Fundamental Services offers the first in a series of stand-alone (SA) ham programs for IBM PCs and compatibles. The SA series has pull-down menus, mouse support, help windows, color displays, and fast assembler subroutines.

"SALOG" is a logger for all bands and modes, the grid locator system, round table QSOs, and net control. Generated logs can be edited and printed.

"SAQSL" prints QSL cards and labels using logs from SALOG. You can print two sizes of continuous form cards and two sizes of labels for QSOs.

The SA series is available from Fundamental Services, 1546H Peaceful Lane N., Clearwater, Florida 34616. The cost is $19.95 each plus $2 shipping and handling; Florida residents add 6-percent sales tax.

Circle #304 on Reader Service Card.

ICOM's New IC-24AT Dual Band Mini Handheld Transceiver

ICOM announces its new 440-MHz and 144-MHz dual band mini handheld, the IC-24AT. Features include:

- Five-watt power output
- Built-in clock
- Crossband full duplex capability
- Forty double-spaced memory channels
- DTMF autodial memory

The IC-24AT also has an external DC power jack, repeater functions, priority watch, and a variety of scan functions. Available options include the UT-50 CTCSS encoder/decoder unit, and UT-51 CTCSS encoder unit.

For details contact ICOM America, Inc., 2380 116th Avenue NE, PO Box C-90029, Bellevue, Washington 98009-9029.

Circle #305 on Reader Service Card.

AEA's New AT-3000 3-kW Antenna Tuner

AEA has a new AT-3000 3-kW antenna tuner, which incorporates the features of the AT-300 300-watt antenna tuner in a high powered package.

Features include:

- 3,000 watts continuous duty cycle
- Front panel switch to select two unbalanced (coax-fed) antennas, a dummy load or a balanced antenna
- Peak and average reading cross-needle meter which shows forward power, reflected power, and SWR

The AEA AT-3000 antenna tuner is available through AEA authorized dealers. For details contact: Advanced Electronic Applications, Inc., 2006 196th Street SW, PO Box 2160, Lynnwood, Washington 98036. Telephone(206)775-7373.

Circle #306 on Reader Service Card.